ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

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CENTRALIZED SUPERVISION FOR A 65-KILOVOLT POWER NETWORK

TELEGRAPH RELAYS WITH RADIO-INTERFERENCE SUPPRESSORS

GAIN EQUALIZATION OF LINEAR SERVOMECHANISMS

COMPANDOR SYSTEM FOR SHORT-HAUL CARRIER TELEPHONY

TRICON INTERLOCK SYSTEM FOR RAILROAD SWITCHING

LOSS FORMULAS FOR SECOND-ORDER HOMOGENEOUS GRADINGS

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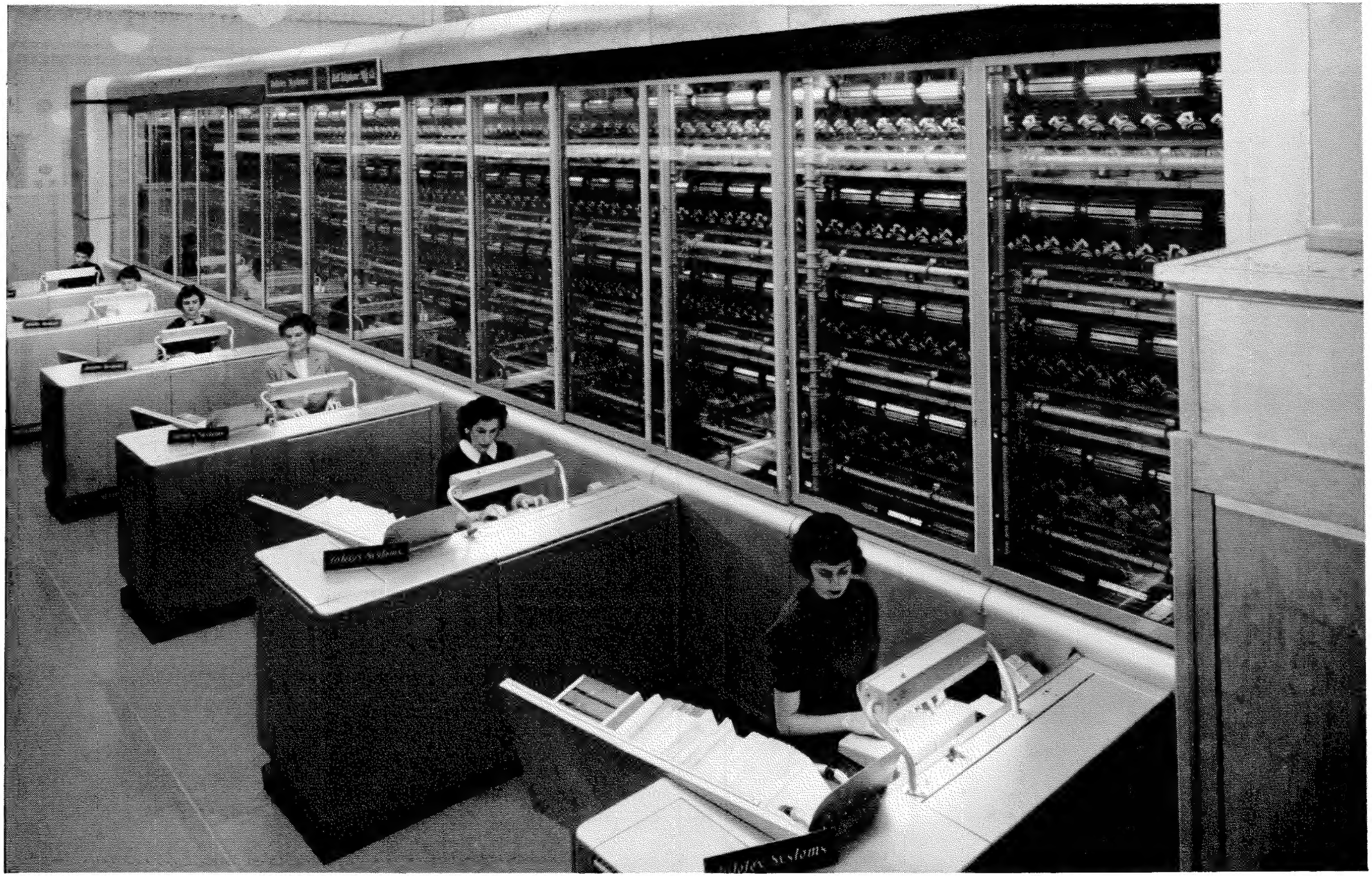

The trend toward automation now includes postal operations. The automatic letter sorter pictured above has been installed in the main post office in Washington, D. C. by Intelex Systems Incorporated; the equipment was designed and manufactured by Bell Telephone Manufacturing Company, Antwerp, Belgium. Operators man six positions of the machine, which can distribute 18 000 pieces of mail per hour to 300 different destination bins. The operators send each letter to its proper destination bin by merely pressing three buttons on a keyboard corresponding to a destination code. The employment of the machine doubles the efficiency of sorting operations over manual sorting methods.

Centralized Supervision of a 65-Kilovolt Power Network

By LUCIEN R. GILLON

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WARTIME destruction stimulated a modernization program of the 65-kilovolt electricity production and distribution system used by several coal mines under control of the Groupe des Houillères du Bassin de Lorraine in the Lorraine region of France. This is the first high-voltage network operated by a European coal or steel pool to be successfully put under remote supervision through the use of telephone-type apparatus.

From 1946 to 1954, coal production was increased from 7 to 13 million tons chiefly as a result of increased mechanization. In addition, plants have been built to produce coke and other coal derivatives that now process nearly a quarter of the coal production. These not only increased substantially the requirements for electric power but existing transformer stations had to be modernized and new ones installed. The major facilities are shown in Figure 1.

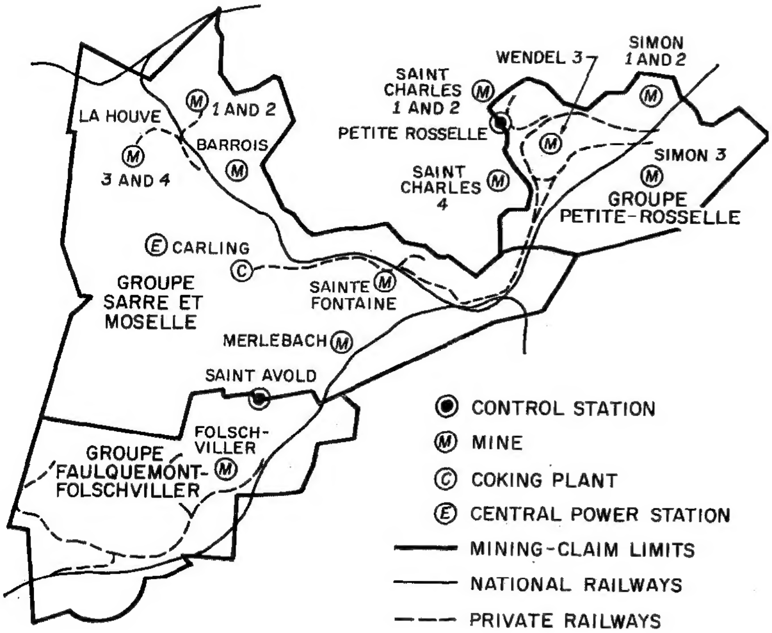


Figure 1—Lorraine coal fields.

The output of these mines and plants is so important that every effort has been made to prevent any power failures. The power equipment and distribution system have been designed to ensure uninterrupted operation and where such might be needed, emergency installations have been provided. Information on the entire system is transmitted to supervisory operating centers at Saint Avold and Petite Rosselle, from which immediate action can be taken to correct any abnormal condition that may arise.

1. Distribution Network

The system takes power from the generating station at Carling and distributes it at 65 kilovolts to 19 transformer stations, the secondary voltages being either 3000 or 5000 volts. There are two control centers, one in Saint Avold for the western sector and the other in Petite Rosselle for the eastern sector. The locations of these control and transformer stations are shown in Figure 2.

The transformer stations are in the vicinity of the operating or ventilating shafts of the mines. As is usual, several stations will be supplied over a loop that permits power to be transmitted in either direction in case of a fault. The few stations not on these loops have duplicate lines to ensure continued operation. Power is also available from the national power network and arrangements are under consideration for a connection to the Sarre system and for duplicate lines between the two control stations. Each transformer station has two sets of incoming busbars to which the transformers may be connected either to ensure operation in case of damage to part of the station or for maintenance work.

2. *Centralized Control*

The design is based on the use of operators at the central control stations to decide and initiate changes that are effected by suitable equipment in the transformer stations. Previously, information was reported by operators in each transformer station to the central control station by telephone and instructions received were put into effect by these operators. Thus, the new system replaces the human operators in the transformer stations with apparatus for both reporting conditions and executing orders.

The reporting of information and the transmission of instructions over substantial distances immediately brings to mind the effectiveness of telephone switching systems for such purposes; the calling of one subscriber by another is an exercise of remote control. The important difference is that of reliability. Wrong telephone calls are not serious unless there are many of them but great damage could result if the wrong transformer station should receive certain orders intended for another station. Safety is therefore the predominant feature underlying the design of remote-control systems for power networks.

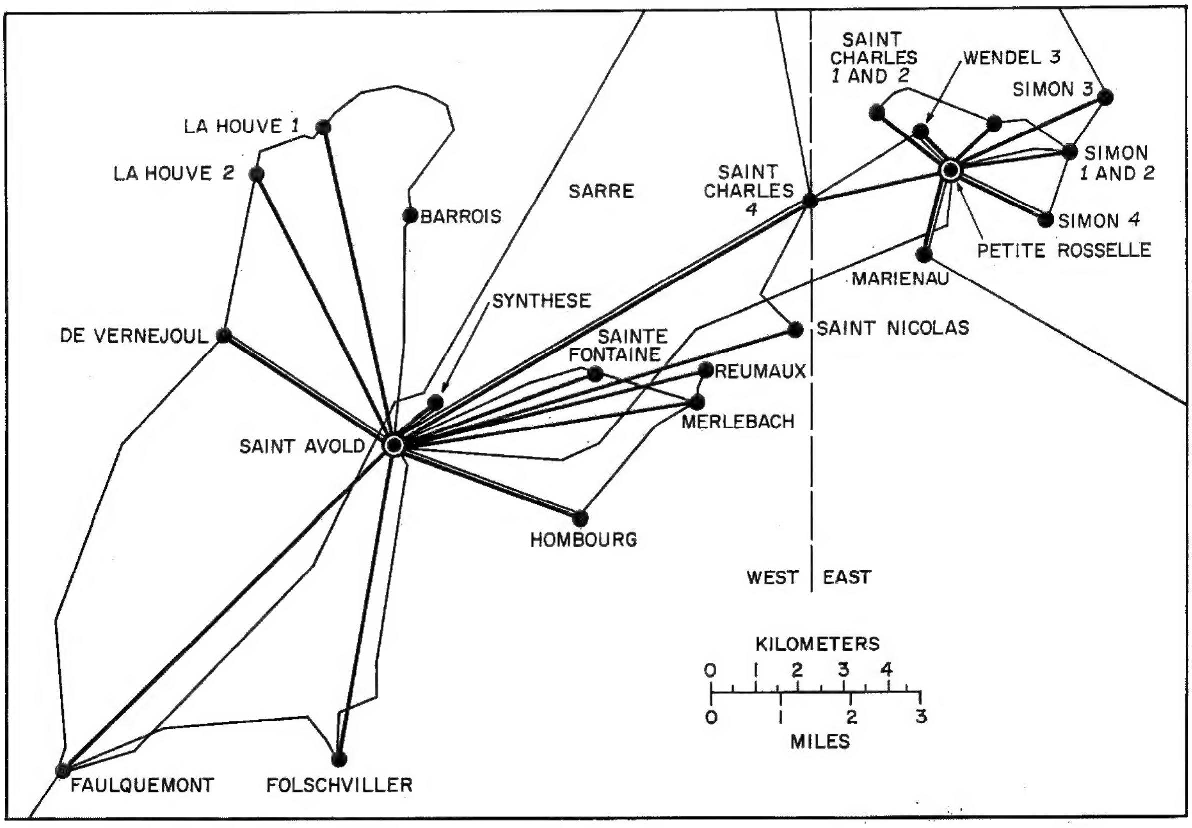


Figure 2—The heavy lines show the direct control connections between each transformer station and its control station. The light lines are the power-distribution circuit.

3. *Reliability*

To insure reliable operation, the design must conform to a number of principles.

Orders for each elementary operation must be confirmed and no action must be taken until the preceding operation has been completed correctly. This permits the system to detect abnormal conditions and to stop any action until the abnormality is corrected.

All apparatus is to be of simple design to avoid malfunctioning. It must have long life, infrequently require servicing, and be capable of operating over wide ranges of input voltages and under varying ambient conditions. The signaling system or code should be simple and errorproof.

The geographical arrangement of the network made it convenient to connect each transformer station to its control station so that signals need not be relayed through tandem stations.

4. *Controlled Elements*

The types of apparatus to be controlled are circuit breakers, section switches, and similar devices that may occupy one of two positions, either *on* or *off*. The positions of these switches must be reported to the control station. In addition, information must also be supplied on faults such as abnormal grounding or overheating and these reports are either *normal* or *fault*. Thus, for control purposes, each of these individual elements is in effect a two-position device.

5. *Operation*

To change the operating condition of any equipment at one of the transformer stations, the operator at the control station, working from the control desk for that particular transformer station, turns a bar switch that is mounted in a mimic diagram and that corresponds to the device to be controlled. A lamp in the center of the bar switch then lights and the operator presses the button, which in turn is mounted on another switch.

The order is transmitted by an individual relay switch associated with the button through a connecting relay to the coding and supervising equipment (Figure 3), which performs the signal transmission.

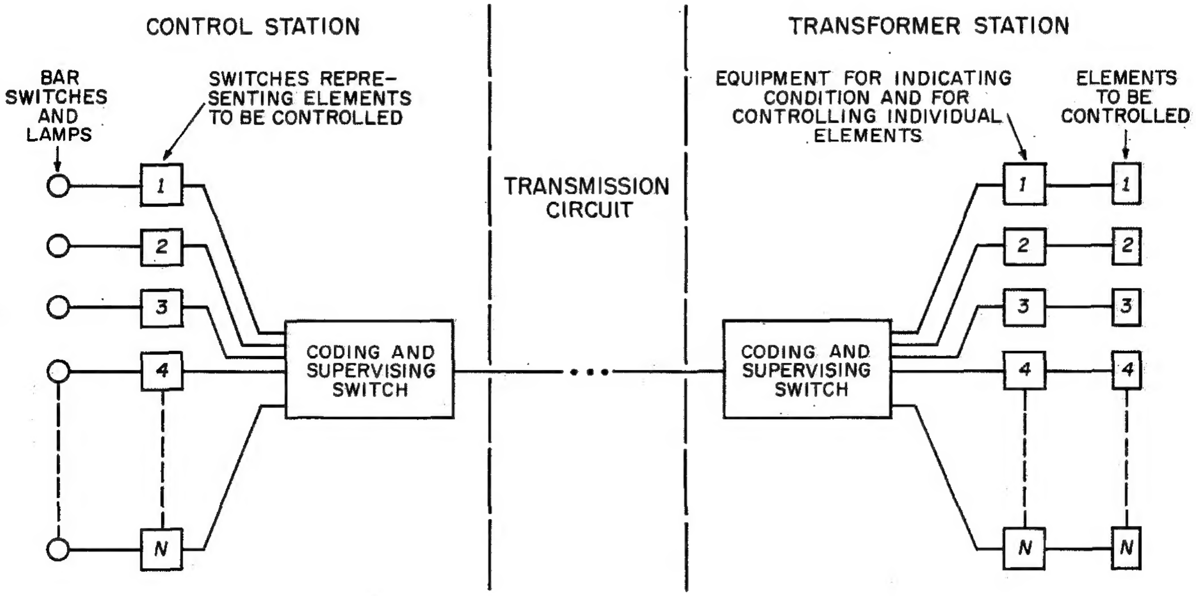


Figure 3—Diagram of controlling and signaling system.

It is therefore required to provide a system that will connect without error of any kind an indicator at the control station with the reporting and controlling mechanisms for the equipment that it represents at the transformer station. Regardless of the number of pieces of equipment that must be controlled at each transformer station, only two pairs of wires for bidirectional signaling are available between that station and the control station. Such an arrangement is shown in Figure 3.

As all the bar switches corresponding to an equipment in a given transformer station can be connected to the coding and supervising switch, which is connected to the transformer station by the transmission circuit, no further transformer-station selection is needed.

A calling code pulse is then automatically transmitted to the finder-selector switch at the transformer station through which the line is connected to the supervisory equipment associated with the individual device to be controlled.

It would now be possible to order the new condition but the correctness of the connection must first be confirmed. This is done by having the connected supervisory equipment transmit its identifying call back to the control station. If this return call corresponds to the original call, the new order will be initiated by the change in the bar switch from its previous position. This transmits the order code to the transformer station and, when the necessary operations have been performed, the new position will be reported by a supervision code to the control station and will extinguish the lamp in the bar switch.

The transformer station will report to the control station any change in conditions that may occur. A control contact on the equipment that has changed condition will initiate a call to the control station and through the coding and supervising switches, each of which actually have both functions, will light the lamp of the corresponding fault indicator on the control board.

6. Code

Information is transmitted by the use of a code in which square-wave pulses are separated by intervals of time approximately equal to the length of the pulses. Pulses are either transmitted or suppressed to distinguish one message or address from all others.

If n is the number of time intervals reserved for pulses and the pulse may assume either of 2 conditions, on or off, the possible number of combinations will be 2^n. In this installation, 10 time intervals are available, so $2^{10} = 1048$ possible combinations. As the message is either of two conditions, the code provides 9 pulse positions for 512 addresses.

The fact that the code consists of a series of pulses and spaces of equal duration permits random disturbances to be rejected because they do not occur at the time intervals set by the code.

6.1 Synchronizing

Each transmitting station sends two signals. One of these is a synchronizing signal A (Figure 4) and the other B contains the address and message.

At the receiving end, the A and B signals must be separated from each other. This can be done in several ways. The A signal may be sent over one pair of wires and the B signal over another pair. This requires a minimum of three wires if earth return is not used. Another method would reverse the polarity of a direct current sent over a single pair of wires, using only one polarity of transmission at a time to avoid the two transmissions canceling each other. A third method would employ alternating currents of different frequencies. The second method is used in this installation.

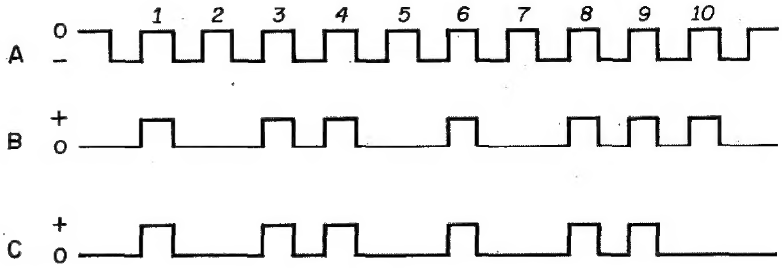


Figure 4—Sample code transmissions. A is the synchronizing signal of negative polarity. B is an address and message signal of positive polarity with the transmitted pulses occurring during the intervals between the A pulses. The first 9 pulse intervals are for the address. C differs from B only in the 10th pulse position, which indicates the condition of or order to the addressed element.

When a call is initiated at the control station, the first pulse of the synchronizing signal A_c is transmitted to the transformer station. It causes the transformer station to reply with its synchronizing signal A_t, the receipt of which at the control station stops the transmission of the first pulse of A_c. As the transformer station no longer receives A_c, it stops its transmission of A_t. There now being no signal in either direction, the cycle is started over again by the transmission of A_c. It is evident that the duration of transmission of each pulse from each station is not rigidly defined but will depend on the time required for the equipments to respond and also on the characteristics of the transmission circuits.

6.2 Address and Message

As already mentioned, the address and message, which make up signal B, are transmitted during 10 pulse intervals that occur between the

synchronizing pulses A. They are distinguished from the synchronizing pulses by being of the opposite polarity. The synchronizing pulses are under the alternate control of both stations, the message of only the sending station.

Each code group is checked to ensure that only one pulse is received during each of the 20 time intervals for the pulses for synchronizing, address, and message. Each group is checked for the proper number of synchronizing pulses. In addition, the address of the transmission must be that of the called device before a message will be accepted. These checks are made for transmissions in either direction between the central control station and a transformer station.

The elapsed time from depressing the push button in the bar switch on the control desk to the extinction of the lamp, which signifies that the order has been filled, but not including the time required to change the condition of the individual element in the transformer station, is between 2 and 3 seconds.

7. *Transmission Circuits*

The maximum distance between a control station and transformer station is 17 kilometers (10.6 miles). This permits the use of 24-volt direct-current signaling.

The power lines, some of which operate at 220 kilovolts, may under fault conditions induce substantial voltages in the signaling circuits, which are in some cases quite close to the power lines. To protect against these conditions, the transmission relays are tested to insure that they will withstand 10 kilovolts between coils, contacts, and ground. The power supplies for the transmission system are insulated to withstand 10 kilovolts to ground. One supply is installed at each central control station. Relays and a power supply may be seen in Figure 5.

8. *Design of Equipment*

The equipment is of the all-relay type. The supervisory apparatus for each individual element in a transformer station is mounted on a removable base that can be replaced by personnel having limited skills. Repairs and adjustments can then be made by trained workers in a repair shop. Each unit is protected by a dust cover.

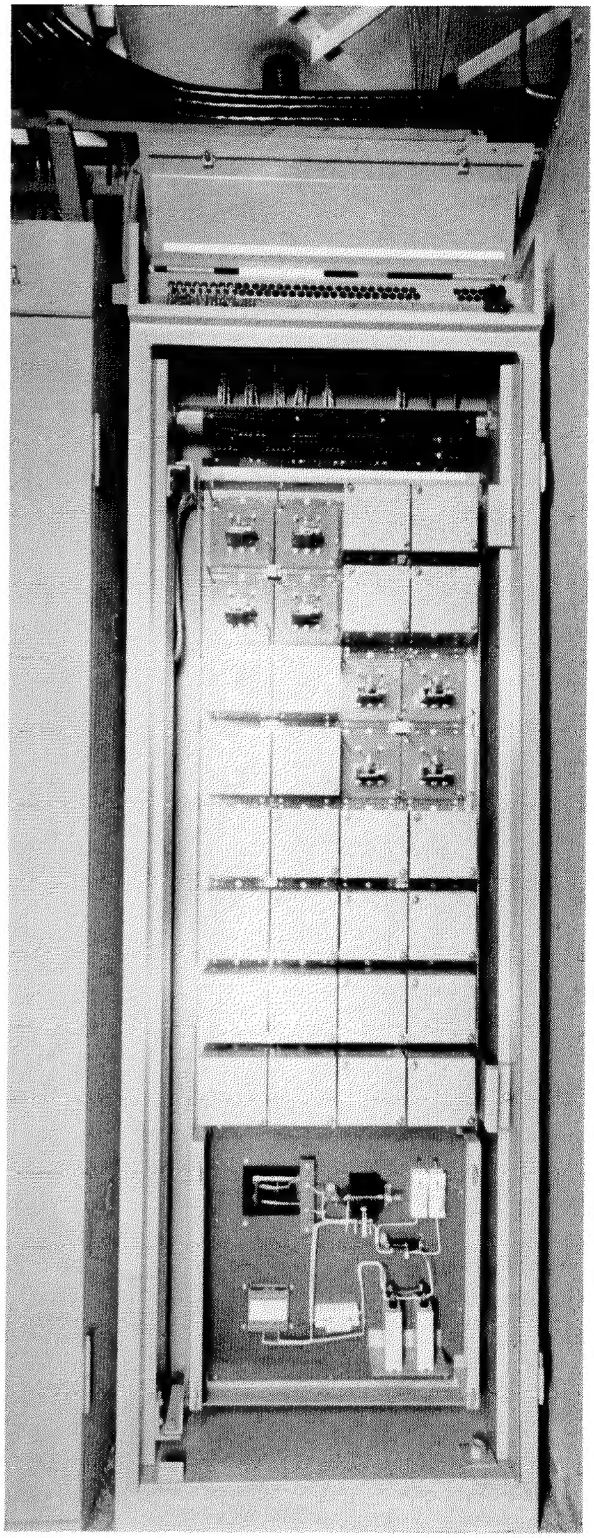


Figure 5—Cabinet mounting signaling relays and power supply, all of which are insulated to withstand 10-kilovolt faults.

8.1 Station Capacities

Table 1 gives the number of individual devices that must be controlled and the number on which reports of condition may be required. It will be noted that Saint Charles *4* is partly controlled from each of the central control stations.

TABLE 1

Individual Elements Requiring Control and Reports of Conditions

Transformer Stations	Individual Elements to be Controlled	Individual Elements on Which Reports Are Made
Synthèse*	24	71
Barrois	7	36
Houve *1*	12	45
Houve *2*	12	42
de Vernejoul*	21	49
Faulquemont*	19	45
Folschviller	10	41
Hombourg	10	41
Sainte Fontaine	12	47
Merlebach*	41	49
Réumaux	26	36
Saint Nicolas	26	36
Saint Charles *4* West	15	38
Saint Charles *4* East*	25	56
Wendel *3**	39	49
Saint Charles *1,2*	14	49
Simon *1,2**	49	71
Simon *3*	10	39
Simon *4*	11	30
Marienau*	39	46

* Maximum capacity is 128 individual elements. All others have maximum capacities of 64 individual elements.

The types of controls and of reports on conditions that may occur in the transformer stations are as follows.

Controls and Position Reports

Line cutoff switches
Busbar cutoff switches
Ground cutoff switches
Line-protecting circuit breakers
Transformer-protecting circuit breakers
Voltage and current measurements
Unlocking of protective devices.

Fault Reports

Line release for current overload
Line release for interphase fault
Line release for phase-to-ground fault
Transformer release for current overload
Transformer high-temperature alarm
Transformer release for overheating
Transformer Buchholz alarm
Transformer Buchholz release
Air-blower failure
Air-compressor failure
Air-compressor overheating
Excessive pressure from air compressor
Inadequate pressure from air compressor
Circuit breaker locked for lack of air pressure
Failure of section direct current
Failure of general direct current
Failure of low-voltage alternating current
Opening of station doors
Local controlled stations
Remote measurement fault
Wiring trouble
Remote-control trouble

Standard equipments are available for 14, 32, 64, and 128 individual elements. Immediate and possible future requirements of transformer stations resulted in the choice of either 64 or 128 elements as shown in Table 1. Figure 6 shows an installation for 128 individual elements. The cabinet at the right will accommodate up to 50 terminal relays; only the number required for the particular transformer station is installed. Others may be added as needed.

8.2 Insulation of Equipment

Telephone-type equipment designed for operation on 24 volts would normally be designed to withstand a test at 500 root-mean-square volts to ground. To permit the use of this equipment in an installation such as this, it is only necessary that the equipment associated directly with the lines subjected to the effects of the power circuits have increased insulation.

There are two points at which the supervisory equipment may be exposed to higher voltages derived from the power system. One is where the supervisory equipment is connected to the circuits that control the power apparatus. The contacts of these relays have been insulated to withstand 2000 volts to ground. The other is where the supervisory equipment is connected to the lines that connect it to the control stations. Here the relay coils and contacts are insulated to withstand 10 kilovolts to ground and to each other.

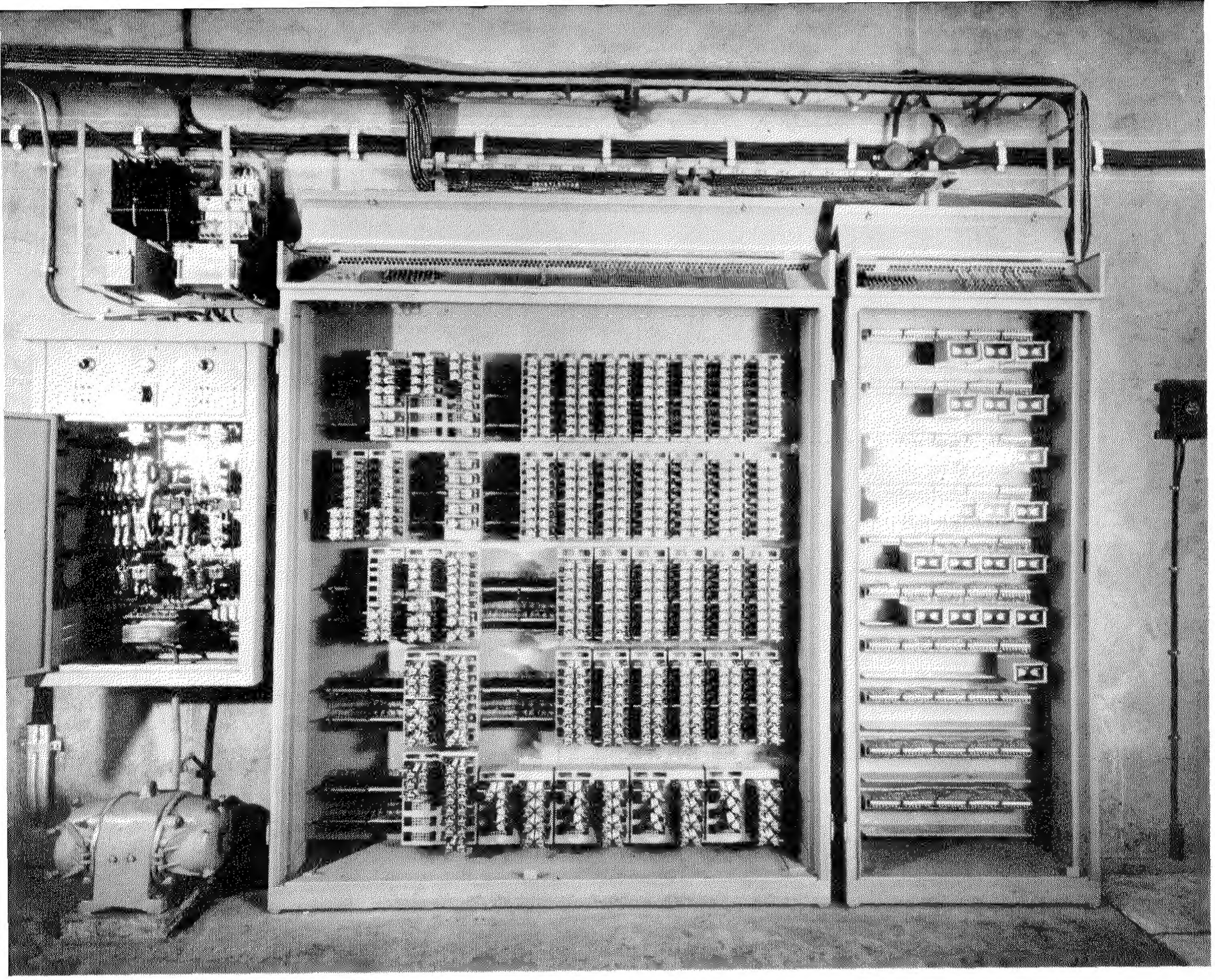


Figure 6—Typical transformer-station installation. The cabinet doors and the covers for the relay groups have been removed. The cabinet at the right will accommodate 50 relays.

8.3 Installation and Testing

To simplify the installation of supervisory equipment, the design permits it to be tested completely within the transformer station and independently of the control station. The connections between the wiring to the controlling mechanisms that operate on the power equipment are made through removable **U**-shaped connectors to the circuits leading to the control station. The removal of the connector will isolate the individual element in the transformer station from all other circuits and permit testing in either direction.

The female contacts into which the **U**-shaped connectors are plugged are arranged in a standard pattern so that a test set may be plugged in to replace the **U** connector. For an individual element that requires reports on its conditions as well as control of its adjusting equipment, the test set provides the following facilities.

A. Check of the continuity of the circuits set up by the control station to open the individual element. This is indicated by two signal lamps.

B. Check of the continuity of the circuits set up by the control stations to close the individual element. This is indicated by two signal lamps.

C. A cutoff and changeover switch on the test set that replaces the contacts of the transformer station power apparatus that reports its condition to the control station. It may be adjusted to report either of the two possible conditions.

Figure 7—Control desk at Petite Rosselle.

All the circuits may be checked with this test set before cutover, which is instituted by inserting the U-shaped connecting links.

In case of trouble, the faulty circuit can be quickly identified by using the test set regardless of whether the fault may be in the transformer station or in the circuits between it and the control station.

9. Control-Station Desks

Great emphasis is placed on reducing the size of the components mounted on the control-station desks so as to permit the operator to have immediate access to the largest number of equipment controls. The desk of the control station at Petite Rosselle is shown in Figure 7.

9.1 Switches and Lamps

Three components have been designed into a miniature unit referred to as a turn-and-push button. It is made up of a position key in the form of a bar switch, control push button, and indicator lamp. The push button has to be large enough to be found quickly and pressed by the operator without danger of its being mistaken for an adjacent unit. Inserted in the mimic diagram of the power system, the bar-switch position key may be turned to indicate the desired condition of the individual element in the transformer station that it represents. The push button mounted in the center of the position key has a Plexiglas window through which the indicator lamp shines.

9.2 Fault Indicators

The fault indicators that identify the various messages are photographic films mounted in Plexiglas frames and illuminated from behind.

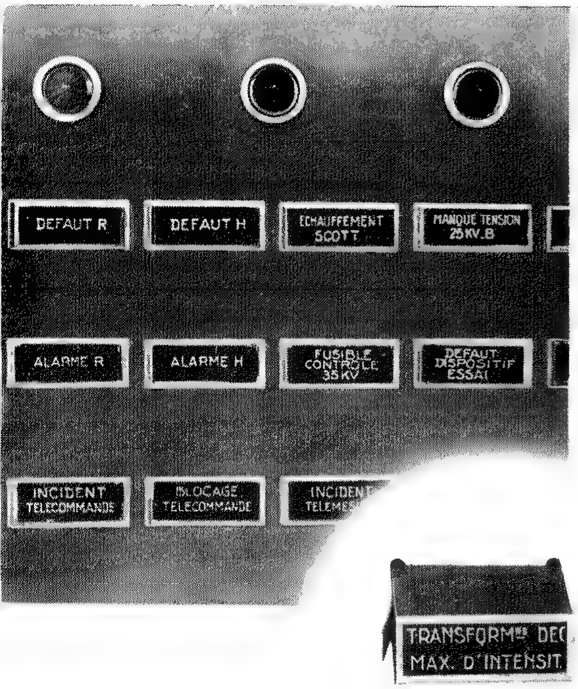


Figure 8—Control-desk fault indicators.

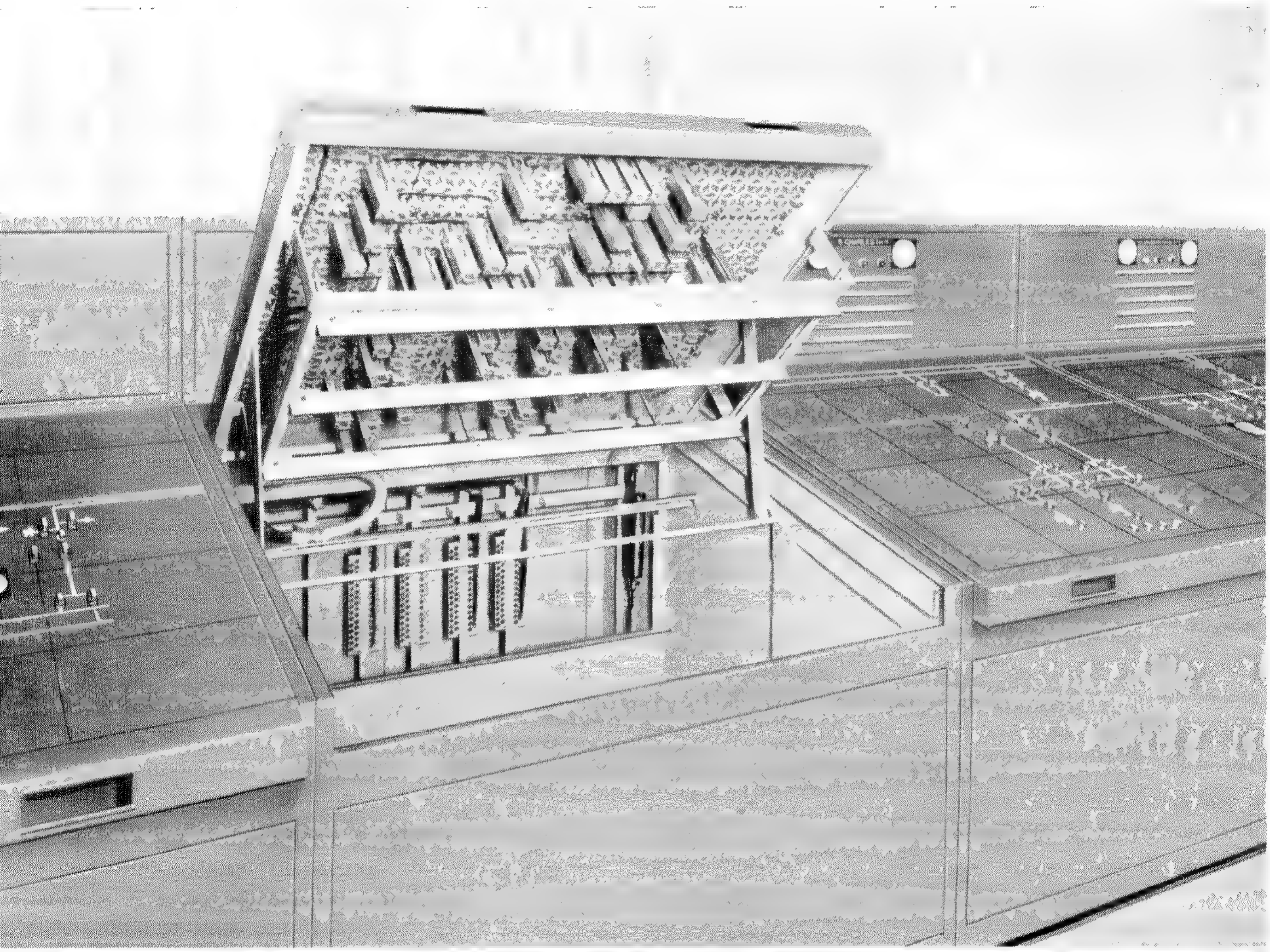
They are shown in Figure 8. The lamps are mounted on removable bars for convenience in replacing them. The films also may be readily replaced.

9.3 Desks

Each control desk is an independent unit that controls one transformer station. The units are mounted next to each other to preserve the continuity of the mimic diagram of that part of the system under the supervision of the particular control station.

The slanting panel of each desk, as shown in Figure 9, mounts the turn- and push-button units for the individual elements of that transformer station. They are arranged as parts of the mimic diagram and are mounted on standard-sized panels to allow for modification and expansion. These 200-millimeter (7.9-inch) square panels are perforated to mount any of the

Figure 9—Typical control desks. Each desk is for one transformer station and the control units are mounted as parts of the mimic diagram.



standard units that make up the mimic diagram. They are not changed when modifying the diagram. Front panels are provided for each type of unit and also carry the corresponding mimic-diagram markings.

The desk top is pivoted and may be held open by springs to permit inspection, servicing, and replacement of indicating lamps. The vertical panel behind the sloping desk top support the fault indicators shown in Figure 8 and the measuring instruments. A removable cover gives access to the back of the desk and to the connecting cables.

10. Recorder

The various types of faults and alarms are reported by the illumination of the corresponding indicator on the vertical panel at the back of the control desk. This lamp is extinguished when the fault is corrected. Records are required of the occurrence and correction of some of these faults and alarms such as those for the air blowers, air compressors, and the opening and closing of the doors of the transformer station.

To avoid the necessity of providing a separate recorder for each device and operating it constantly, an arrangement has been developed whereby any fault to be recorded sets a common mechanism in operation. The recorder hunts over the circuits until it finds the one indicating a fault, notes which circuit it is, the time the fault appeared, and the time it disappeared. This information is recorded on tape by a standard teleprinter of the type shown in Figure 10.

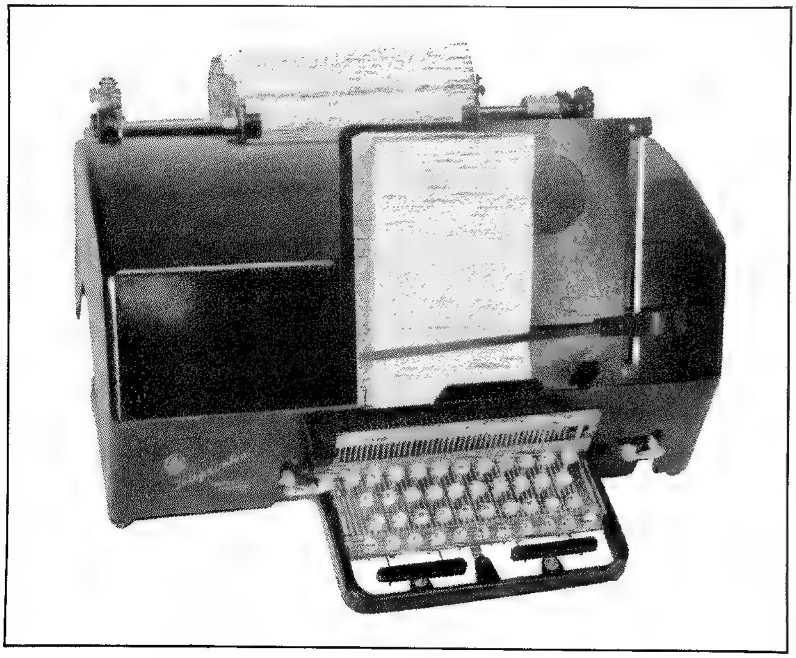


Figure 10—Teleprinter used for recording faults.

11. Telemetering

It is important for the control operation to know under certain conditions such things as the voltages, active and reactive powers on the transformer-station lines, and current supplied by the transformers. These values may be requested by the operator and will be transmitted to the control station and indicated on the measuring instruments mounted on the vertical panel with the fault indicators.

12. Conclusion

This installation uses about 10 000 relays for the supervision of some 1300 individual elements in the 20 transformer stations. Normally, each transformer station requires a team of 3 operators or 60 persons in all. A central control station with a mimic diagram and connecting telephone lines to each transformer station would also be required.

Remote control has eliminated the need for transformer-station personnel; a crew of 4 control operators is able to supervise the low-voltage as well as the high-voltage systems. The supervisory methods are such that they could be extended to many other equipments, such as pumps, ventilators, and elevators and would require connecting lines equivalent only to poor-quality telephone channels.

The use of telephone-type equipment, the reliability of which has been well established, permits great flexibility in design, small size, low power consumption, and operation over distances of several hundreds of kilometers to provide full supervision of a great variety of equipment.

Telegraph Relays with Built-In Radio-Interference Suppressors

AS IS WELL KNOWN, the square-wave voltage normally transmitted by the armature of a polarized relay reproducing telegraph signals is rich in harmonics extending well into the radio-frequency range. For example, the Fourier analysis of a 25-cycle-per-second square-wave signal of 80 volts peak (that is, a continuously repeated polar signal of alternate 1-unit marks and 1-unit spaces at 50 bauds from a telegraph battery of 80–0–80 volts), shows that this signal contains voltages of approximately 85 to 5000 microvolts in the frequency band 30 megacycles to 500 kilocycles per second. This is quite an appreciable signal strength to a radio receiver and consequently, when telegraph and radio equipments are used in close proximity, it is desirable to suppress the radio frequencies without materially deteriorating the effective squareness of the signal as far as the telegraph equipment is concerned.

The basic circuit for such a suppressor (developed by Standard Telephones and Cables, Limited, of London, England) is a low-pass filter and the circuit used is shown in Figure 1; it is intended primarily for relays transmitting polar (double-current) signals. The resistors *R1* and *R2* are to reduce the surge currents that occur when capacitors *C1* and *C2* discharge through the relay contacts and which would otherwise cause contact sparking. It should be stressed that the suppressor supplements the normal spark-quench circuit and is not intended to replace it. The spark-quench circuit itself will give some radio-interference suppression, chiefly at the lower end of the frequency spectrum. The coils are "Carbonyl" dust-cored toroids and are capable of withstanding the full fault current should a ground or a short-circuit occur.

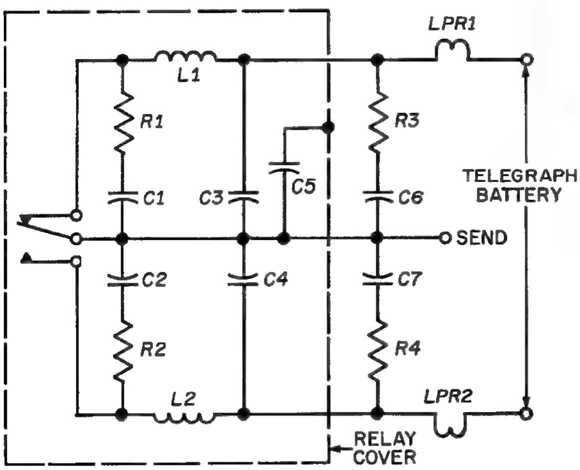
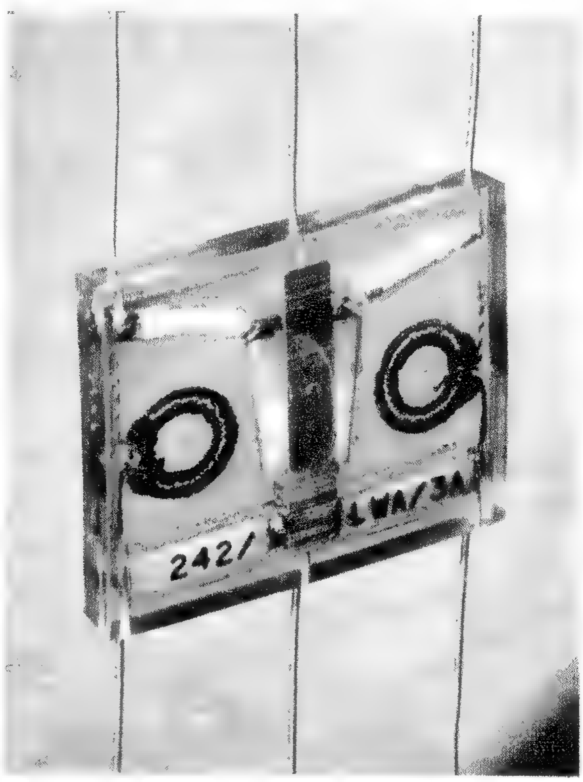
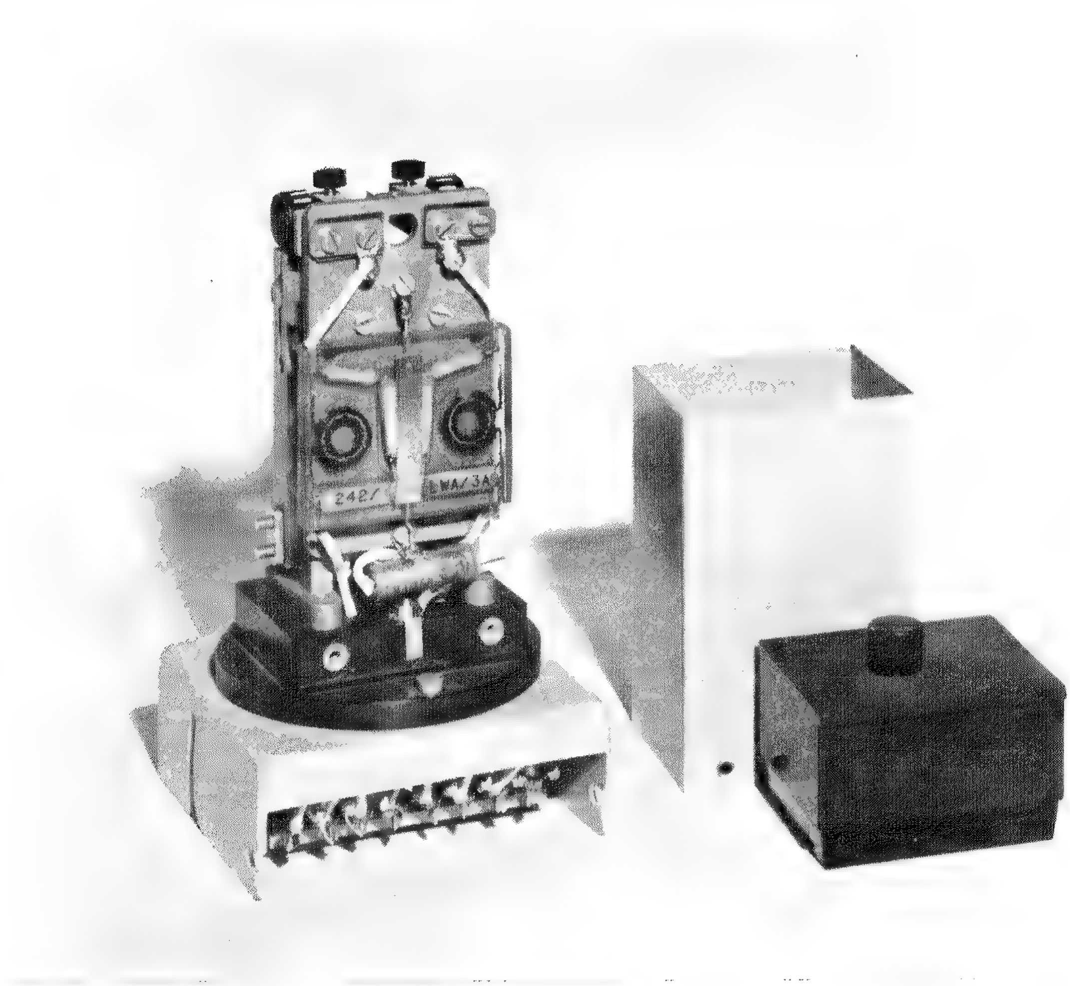


Figure 1—Schematic diagram of the interference suppressor. *R1* = *R2* = 1000 ohms. *C1* through *C5* = 0.01 microfarad. *L1* = *L2* = 1 millihenry. The usual spark-suppressor circuit consists of *R3* = *R4* = 1500 ohms and *C6* = *C7* = 0.5 microfarad. *LPR1* and *LPR2* are protective lamps.



Type-*242/LWA/3A* suppressor.

The components for the suppressor unit are encapsulated in "Stantelene *V*," which is a plasticized expoxy-type resin, so as to form a rigid assembly whose overall dimensions can be controlled. The assembly can then be mounted and wired within the relay cover right up against the interfering source, just where it ought to be.



Type-*4191* relay fitted with *242/LWA/3A* suppressor.

Two forms of suppressor are made; one for a "Standard" *4191*-type relay in which all components for the suppressor are moulded into one unit (coded *242-LWA-3A*) and mounted at the back of the relay framework; the other for the "Standard" *4192*-type relay in which space is more restricted and the components are therefore moulded into two equal units (coded *242-LWA-4A*), one unit being mounted on each side of the relay framework (the 5th capacitor is wired on separately). With each type of suppressor, the units are within the relay covers and the rigidity of the mouldings and of the wires is sufficient to keep them in position without having to add fixing holes and screws.

These suppressors, when assembled within the relay framework, give a suppression of at least 50 decibels in the range 0.5 to 30 megacycles per second *when measured directly at the base of the relay;* panel wiring and panel dust-covers will give a further improvement in suppression.

The sizes and weights of the suppressor units are:

	Length		Width		Thickness		Weight	
	Inches	Centimeters	Inches	Centimeters	Inches	Centimeters	Ounces	Grams
242-LWA-3A	1.562	3.96	1.875	4.76	0.265	0.67	0.780	22.2
242-LWA-4A	1.375	3.49	0.875	2.22	0.25	0.63	0.3	8.7

Gain Equalization of Linear Servomechanisms that Solve Nonlinear Equations

By GUY E. ADAMS

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LINEAR servomechanisms are frequently used to solve nonlinear equations. Gain equalization of the servomechanism loop may be required to achieve adequate accuracy and stability. This paper presents a method wherein the error equation is differentiated with respect to the output rotation to determine the required gain equalization function. Several examples of the application of this method are given. The method is also applied to a multiple-loop system that solves three nonlinear equations.

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Servomechanisms are used to solve linear equations in most of the applications. In such cases, it is desired that the input and the controlled output be equal or vary according to a constant relation; that is, the steady-state output may be described by multiplying the input by a constant. The basic open-loop servomechanism is considered to be linear.

Servomechanisms may also be used to solve continuous single-valued nonlinear equations. The closed-loop servomechanism becomes nonlinear even though the basic open-loop system is linear; in this paper the open-loop system is treated as being linear and damped sufficiently for stability at the nominal gain-equalized sensitivity. The servomechanisms referenced by this paper may contain the nonlinear element as any part of the loop; however, in equation-solving systems, the computing element may be considered to be either the error-sensing device or part of the feedback path. Gain-equalization methods described by this paper apply equally well when the nonlinear device is part of the forward path; however, the forward-path location generally corresponds to an undesirable system characteristic.

It will be shown that the nonlinear-equation-solving servomechanism has variable loop gain, which may result in inconsistent or variable system performance. If the amplifier gain is adjusted for some average value, the loop gain at one extreme of the range may be too high and the gain at the other too low. The well-known effects of too-high gain are various forms of instability such as:

A. Oscillation or limit cycles.

B. Gear chatter.

C. Hum caused by resolution limitations in the follow-up member.

D. Excessive amplification of noise and disturbance signals.

E. Excessive power consumption.

Too-low gain may result in sluggishness and inaccurate follow-up. Conditionally stable servomechanisms are especially susceptible to gain variations.

1. Theory of Gain Equalization

Gain equalization is a solution for the undesirable effects of variable loop gain. The method described in this paper utilizes the derivative with respect to the output of the input–output-error equation to equalize the servomechanism loop gain. The gain is equalized by a controller gain function that is approximately proportional to the reciprocal $1/(d\epsilon/d\theta_o)$ of the error sensitivity.

The gain error in using the reciprocal error sensitivity function is generally small in practical cases because the variations in output position of the servomechanism about the required position are small. The derivative $d\epsilon/d\theta_o$ may be approximated accurately by the ratio $\Delta\epsilon/\Delta\theta_o$ in such cases.

The amount of gain variation that is tolerable without gain equalization is beyond the scope of this paper. Generally, gain equalization would be used only where loop-gain variations were fairly high, say 10 decibels or more.

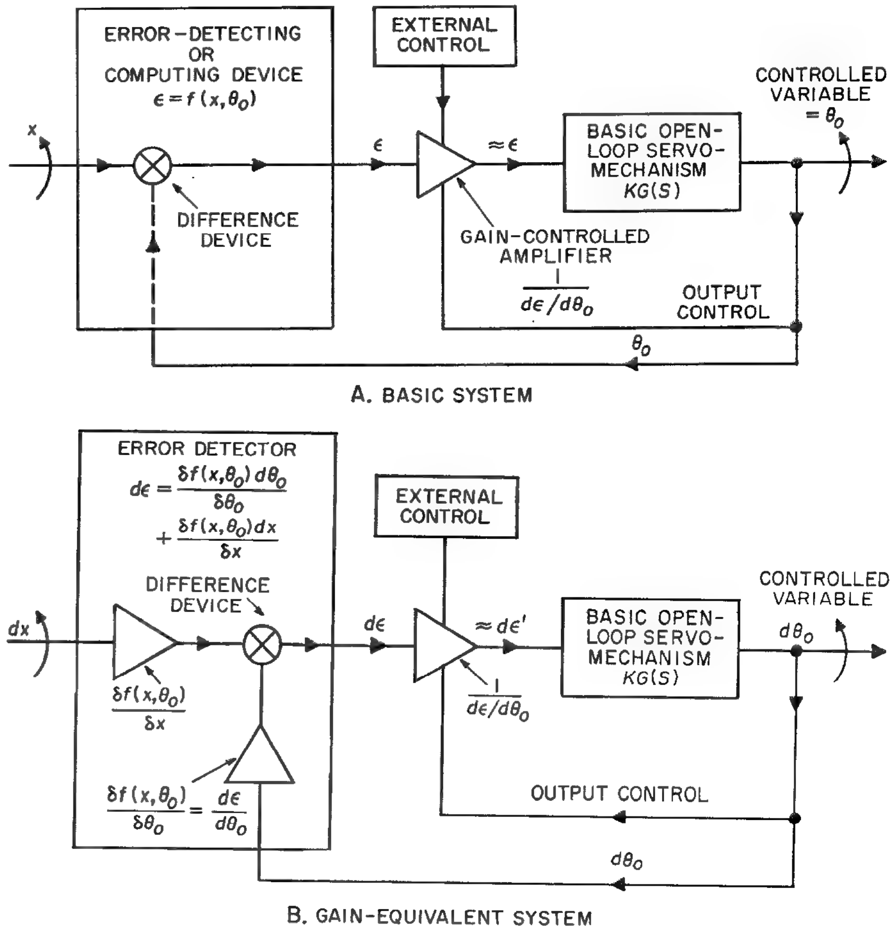


Figure 1—Servomechanism with gain equalizer for solving nonlinear equations.

For A,

$R \gg R_p$

$\epsilon = E - x\theta_o$

= error-detector equation

$\theta_o = \left.\frac{E}{x}\right|_{\epsilon=0}$

= computer equation

$|d\epsilon'/d\theta_o| = EA$

= gain-equalized error-detector sensitivity.

For B,

$$\frac{\epsilon'}{\epsilon/2} = +\frac{A_2}{1+A_2x}$$

$$\doteq +\frac{1}{x} \text{ when } A_2x \gg 1$$

$$\frac{d\epsilon'}{d\theta_o} = \frac{d\epsilon}{d\theta_o}\cdot\frac{\epsilon'}{\epsilon/2} \doteq 1.$$

The basic scheme of nonlinear-equation solving and gain equalization is illustrated by the servomechanisms of Figure 1. The error detector in Figure 1A produces an error ϵ that is a desired function of the input variable x and the controlled variable θ_o. A gain-controlled amplifier accepts the error voltage ϵ and produces a gain-equalized error voltage ϵ' that is a linear function of θ_o. The equalized error voltage is applied to the input of an ordinary open-loop system. The gain-controlled amplifier has a gain proportional to $1/(d\epsilon/d\theta_o)$. The gain might be controlled by the output position, or by an external means that is some function of the input variable, or by both of these.

The gain equivalent circuit of Figure 1A is shown in Figure 1B, which represents the system for small values of error, small changes in input, and small changes in output. This circuit may be obtained by using the following derivation: Assume that (1) in functional notation describes the input, output, and error relations of the system.

$$\epsilon = f(x,\theta_o). \tag{1}$$

Differentiating (1),

$$d\epsilon = \frac{\delta f(x,\theta_o)}{\delta\theta_o}\,d\theta_o + \frac{\delta f(x,\theta_o)}{\delta x}\,dx. \tag{2}$$

The differential error relation can be represented by the error detector in Figure 1B. The differential input dx is amplified by $\delta f(x,\theta_o)/\delta x$ in a fictitious amplifier and is then applied to a difference device. The incremental output, $d\theta_o$

is amplified by $\delta f(x,\theta_o)/\delta\theta_o$ in a second fictitious amplifier and is applied to the same difference device. The differential error $d\epsilon$ is applied to the gain-equalizing amplifier. The error sensitivity is $\delta f(x,\theta_o)/\delta\theta_o$ and is independent of the magnitude of dx. The equation for $d\epsilon'$, the error with gain equalization, is

$$\left.\begin{aligned} d\epsilon' &= \frac{d\epsilon}{d\epsilon/d\theta_o} \\ &= d\theta_o + \frac{[\delta f(x,\theta_o)/\delta x]dx}{\delta f(x,\theta_o)/\delta\theta_o}. \end{aligned}\right\} \quad (3)$$

Equation (3) shows that the gain has been equalized because $d\epsilon'/d\theta_o = 1$ with $dx = 0$; it also shows that the change in output $d\theta_o$ as a result of an input dx is the same as without the equalizing network.

The remainder of this paper is devoted to demonstrating how this method of gain equalization has been applied to several different servomechanisms that solve nonlinear equations, so that the rather-wide range of application from simple to moderately difficult problems can be demonstrated.

2. *Reciprocal Computor*

A diagram of a gain-equalized servomechanism to provide a rotation that is the reciprocal of the input command is shown in Figure 2. The equation for the error ϵ is

$$\epsilon = E - x\theta_o. \quad (4)$$

The equation is solved when $\epsilon = 0$; the output rotation is the reciprocal of the input voltage:

$$\theta_o = - E/x. \quad (5)$$

The error sensitivity may be found by differentiating (4) with respect to θ_o.

$$d\epsilon/d\theta_o = - [x + \theta_o(dx/d\theta_o)]. \quad (6)$$

Since the input is not a function of the output rotation, (6) simplifies to

$$d\epsilon/d\theta_o = - x. \quad (7)$$

The desired relation for error sensitivity is obtained by substituting the value for x from (5):

$$d\epsilon/d\theta_o = - E/\theta_o. \quad (8)$$

The required gain-equalizing function becomes

$$1/(d\epsilon/d\theta_o) = - \theta_o/E. \quad (9)$$

The gain-equalizing circuit consists of an isolation amplifier having a gain of $2A$ and a linear voltage divider positioned by the output shaft. The voltage-divider output is a voltage proportional to θ_o as required by (9). The over-all gain of the error detector and the gain-equalizing circuit for a small variation in output is

$$d\epsilon'/d\theta_o = (- E/\theta_o)(A_1\theta_o) = - EA_1, \quad (10)$$

which is the product of the error-detector sensitivity and the gain of the equalizing circuit. Equation (10) shows that the loop gain has been linearized since an error variation and an output variation differ only by a multiplying constant.

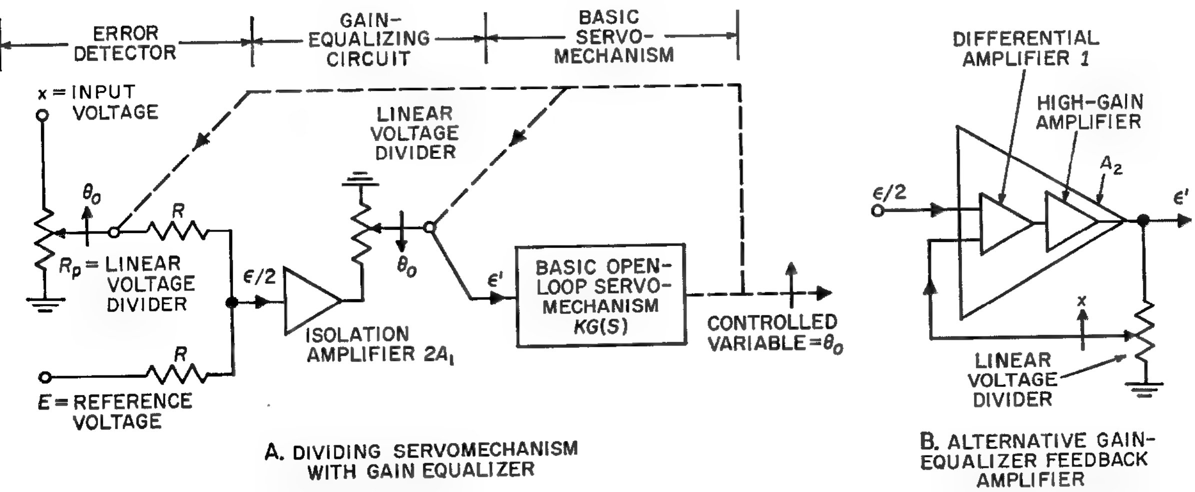


Figure 2—Gain-equalized servomechanism for determination of reciprocals.

If a certain value of combined error-detector and gain-equalizer sensitivity is desired, it may be calculated using (10): In this case, the required A_1 is

$$A_1 = (d\epsilon'/d\theta_o)/E. \tag{11}$$

If a voltage-divider rotation proportional to the input x is available, the sensitivity of the gain-equalizing circuit can be controlled as a function of the input. Referring to (7), the gain can be controlled by a factor proportional to $1/x$ and the alternative gain-equalizing circuit of Figure 2B (a high-gain feedback amplifier) may be used.

3. *Arc-Tangent Computer*

A second example of gain equalization is used in a servomechanism that provides a rotation proportional to $\tan^{-1}(y/x)$, Figure 3. The equation for the error voltage is

$$\epsilon = E(y\cos\theta_o - x\sin\theta_o), \tag{12}$$

where x and y are fractional parts of E.

This equation can be solved for the output rotation θ_o when the error becomes zero.

$$\theta_o = \tan^{-1}(y/x). \tag{13}$$

The error sensitivity may be found by differentiating (12). The sensitivity becomes, treating x and y as constants,

$$d\epsilon/d\theta_o = -Ey\sin\theta_o - Ex\cos\theta_o. \tag{14}$$

By the use of a trignometric identity, (14) may be written as

$$d\epsilon/d\theta_o = -Ex/\cos\theta_o. \tag{15}$$

The gain required from the equalizing circuit is proportional to

$$1/(d\epsilon/d\theta_o) = -(\cos\theta_o)/Ex. \tag{16}$$

The gain-equalizing amplifier of Figure 3 has the required gain function. An amplifier having a feedback factor proportional to x, which is obtained by using a linear voltage divider and an x rotation, provides a gain at its output of $1/x$. The output of the amplifier is applied to a cosine voltage divider positioned by the servomechanism output shaft. The gain of the entire equalizing circuit is the quantity $-(\cos\theta_o)/x$.

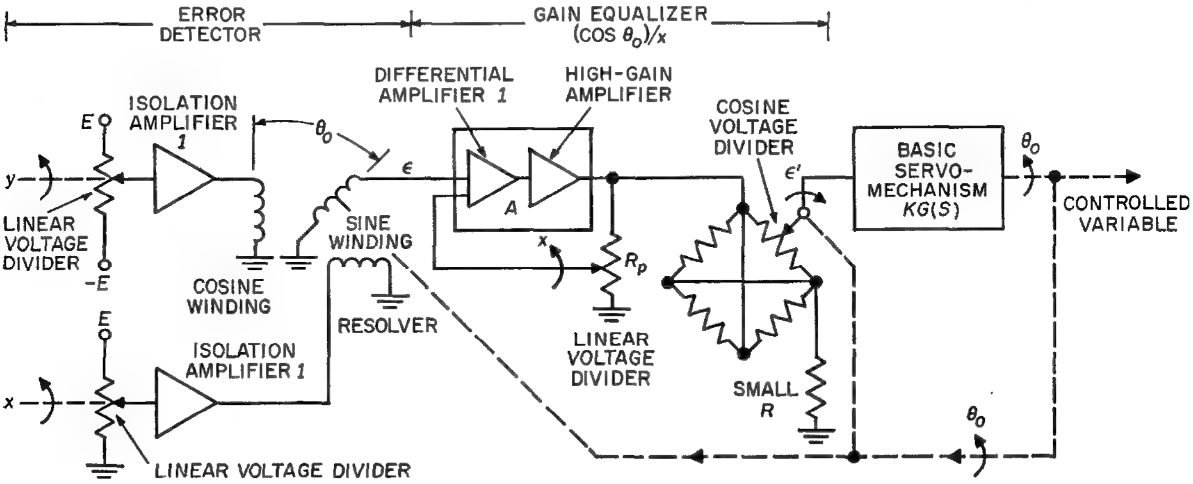


Figure 3—Gain equalization of servomechanism for computing $\tan^{-1}(y/x)$.

$$\epsilon = Ey\cos\theta_o - Ex\sin\theta_o \text{ and } \theta_o = \tan^{-1}\frac{y}{x}\bigg|_{\epsilon=0}\bigg]_{-90}^{+90}.$$

The circuit and servomechanism have several limitations. The maximum gain of the feedback amplifier is A, although infinite gain is required in (16) for $x = 0$. A small resistor can be used in series with the cosine voltage divider to prevent the gain from becoming zero at the ± 90-degree positions; the servomechanism could become

"stuck" at these positions if precautions were not taken.

It can be shown that (13) also can be written

$$d\epsilon/d\theta_o = E(x^2 + y^2)^{1/2}. \qquad (17)$$

A voltage proportional to this quantity can be obtained from the output winding on the resolver in quadrature with the error-voltage winding. The alternating voltage so obtained can be applied to control the output of the power-supply circuit, which furnishes the voltage-divider voltage E. Voltage E is controlled so that the quantity $E(x^2 + y^2)^{1/2}$ from the resolver remains constant; the principle of operation is similar to automatic gain control. This circuit has the disadvantage that the error sensitivity ϵ would generally be considerably less than that of Figure 3 because of power output limitations of

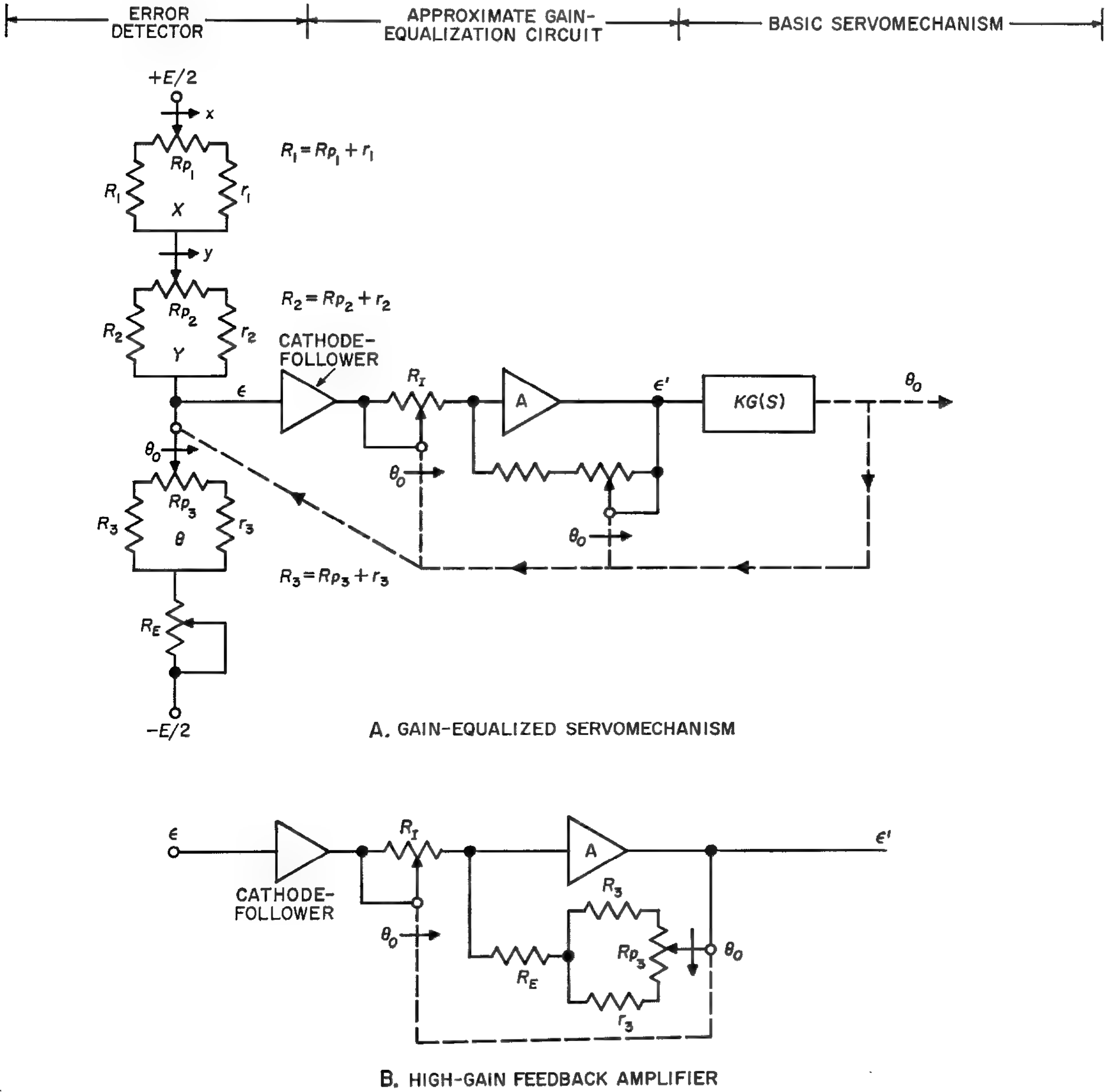


Figure 4—Diagram of square-root computer. In B,

$$\frac{\epsilon'}{\epsilon} = \frac{R_E + R_3/2 - \theta_o^2/2}{\theta_o R_1} \quad \text{where} \quad R_I\theta_o \gg \frac{R_I\theta_o + R_E + R_3/2 - \theta_o^2/2}{A}.$$

the gain-controlled power supply and power dissipation limitations of the X and Y input voltage dividers. It can be shown that the error voltage sensitivity at the equalized output in Figure 3 is higher by a factor $1/(x^2+y^2)^{1/2}{}_{\min}$ than that using the gain-controlled alternating-current power supply.

4. Square-Root Computer

The third example of gain equalization is that required by a servomechanism providing a rotation proportional to the square root of the sum of the squares of the input variables. The system of Figure 4 extracts the square root of the sum of two squares.

The impedance of the X element is

$$Z_x = \frac{(R_1+x)(R_{p1}+r_1-x)}{(R_1+R_{p1}+r_1)}. \qquad (18)$$

Let

$$R_1 = R_{p1} + x; \qquad (19)$$

then,

$$\left.\begin{aligned} Z_x &= (R_1^2 - x^2)/2R_1 \\ &= R_1/2 - x^2/2R_1. \end{aligned}\right\} (20)$$

If the corresponding relation is true in the Y and θ elements, the error voltage ϵ is

$$\epsilon = -\frac{E(Z_x+Z_y-Z_\theta-R_E)}{Z_x+Z_y+Z_\theta+R_E}. \qquad (21)$$

In (20), it may be seen that a typical element Z_x consists of a constant resistance and a variable resistance. Assuming x, y, θ, and ϵ are zero, the constant resistances must have the relation

$$R_E + R_3/2 = R_1/2 + R_2/2. \qquad (22)$$

Substituting in (21), the error becomes

$$\epsilon = -\frac{E(Z_x+Z_y-R_E-R_3/2+\theta_o^2/2R_3)}{Z_x+Z_y+R_E+R_3/2-\theta_o^2/2R_3}. \qquad (23)$$

The error sensitivity is

$$\left.\begin{aligned} \frac{d\epsilon}{d\theta_o} &= -\frac{E[(Z_x+Z_y+R_E+R_3/2-\theta_0^2/2R_3)(\theta_0/R_3)+(Z_x+Z_y-R_E-R_3/2+\theta_0^2/2R_3)(\theta_0/R_3)]}{(Z_x+Z_y+R_E+R_3/2-\theta_0^2/2R_3)^2} \\ &= -\frac{2E(\theta_0/R_3)(Z_x+Z_y)}{(Z_x+Z_y+R_E+R_3/2-\theta_0^2/2R_3)^2}. \end{aligned}\right\} (24)$$

Recalling that at balance,

$$Z_x + Z_y = R_E + R_3/2 - \theta_o^2/2R_3, \qquad (25)$$

the error sensitivity simplifies to

$$\frac{d\epsilon}{d\theta_0} = -\frac{(E/2)\theta_o/R_3}{R_E+R_3/2-\theta_o^2/2R_3}. \qquad (26)$$

Equation (23) is an approximation because of the balance condition assumed in obtaining (26) from (24), but it is sufficiently accurate for gain-equalization purposes.

The equation shows that the error sensitivity is directly proportional to θ_o and inversely proportional to a parabolic function of θ_o. The sensitivity becomes zero for $\theta_o = 0$. The variation in the denominator may be comparatively small, because θ_o is always less than R_3.

The inverse function, (27), is needed for gain equalization.

$$\frac{1}{d\epsilon/d\theta_0} = -\frac{R_E+R_3/2-\theta_o^2/2R_3}{2E\theta_0/R_3}. \qquad (27)$$

The high-gain feedback amplifier circuit of Figure 4B provides a close approximation to (27). A feedback resistance identical with the θ and R_E elements of the error detector can be used to supply the term in the numerator. A cathode-follower and a voltage divider having a resistance proportional to θ_o are the input elements. The cathode-follower is needed to provide a low output impedance for R_I; the additional series resistance in the form of amplifier output impedence is a source of inaccuracy.

The approximate gain-equalization circuit shown in Figure 4A provides circuit simplification by linearizing the numerator. Linearization as used in one application caused a maximum gain error of about -2.5 decibels. In this case, relative circuit values were

$$R_3 = 1.07\ R_{p3}$$
$$r_3 = 0.07\ R_{p3}$$
$$R_E = 0.225\ R_{p3}.$$

Linearization was accomplished by fitting to the end points of the numerator function. The required gain equalization becomes

$$\left.\begin{aligned}\frac{1}{d\epsilon/d\theta_o} &= -\frac{2R_{p3}(0.225+0.535-\theta_o^2/2.14)}{E\theta_o/1.07}\\ &\doteq -\frac{2R_{p3}(0.76-0.466\theta_o)}{E\theta_o/1.07}.\end{aligned}\right\} \quad (28)$$

The curves shown in Figure 5 show the theoretically exact and the approximate values of gain equalization required. The dotted line indicates the measured divergence of the equalizing circuit from the theoretically approximate curve using an amplifier having an open-loop gain of about 75 decibels. In this case, the gain was limited by the output impedance of the cathode-follower. The maximum gain ratio (maximum gain/minimum gain) of the closed-loop amplifier was 200:1.

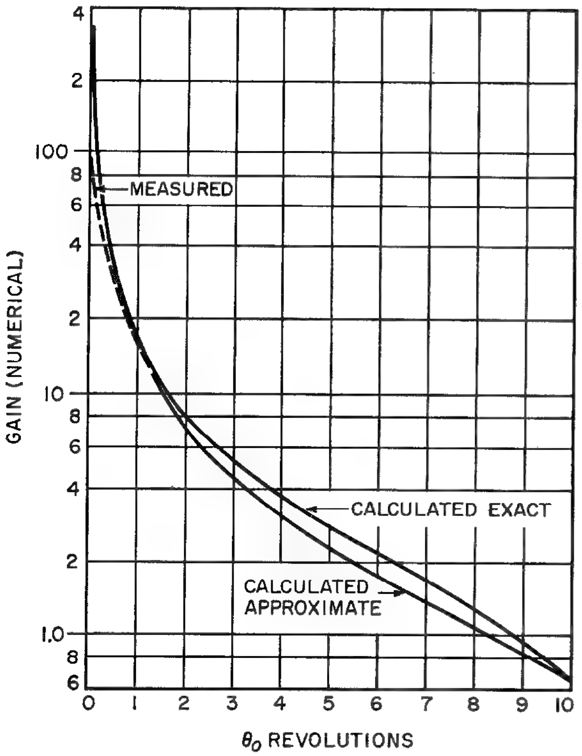


Figure 5—Gain-equalization curve.

Another method that has been used for gain equalization of the computer in Figure 4 is to control the magnitude of the computer supply voltage. A characteristic of this method is that the computer supply voltage (having the same form as Figure 5) is required to be small over most of the range of operation to prevent excessive power dissipation in the computer elements near the zero end of the range. Several disadvantages result:

A. Low error sensitivity requires a higher-gain amplifier.

B. Complicated balanced-to-ground computer power supply circuits are required.

C. Power-supply balance poses a problem, particularly during high slewing rates near the zero end of the range.

5. Discussions on Simple Systems

The method outlined for determining error-sensitivity and gain-equalization requirements generally produces satisfactory results. In each of the three foregoing servomechanism computers, the gain-equalization circuits produced the desired results; that is, frequency response, degree of stability, and angular accuracy were found to remain constant over the range of operation except as noted where extremely high gain was required from the gain-equalization circuit. In reviewing the method of determining error sensitivity and of using the reciprocal error-sensitivity function for gain equalization, one sees that the function $d\theta_o/d\epsilon'$ in Figure 1B is a constant; therefore, positional accuracy and stability are constant. On the other hand, if gain equalization is not applied, the positional accuracy, assuming a small amount of static friction, is proportional to the reciprocal of the error sensitivity.

These methods have been applied successfully to a variety of simple servomechanism computers that solved trigonometric, logarithmic, and algebraic equations. In addition, a multiple-loop nonlinear computer has been gain equalized, but with less-satisfactory results. The principal problem in gain equalizing a multiple-loop nonlinear servomechanism is that equalization can be applied as a general rule only to the error-voltage path of each system and not to principal data-transmission paths; therefore, other expedients may have to be employed to secure stability.

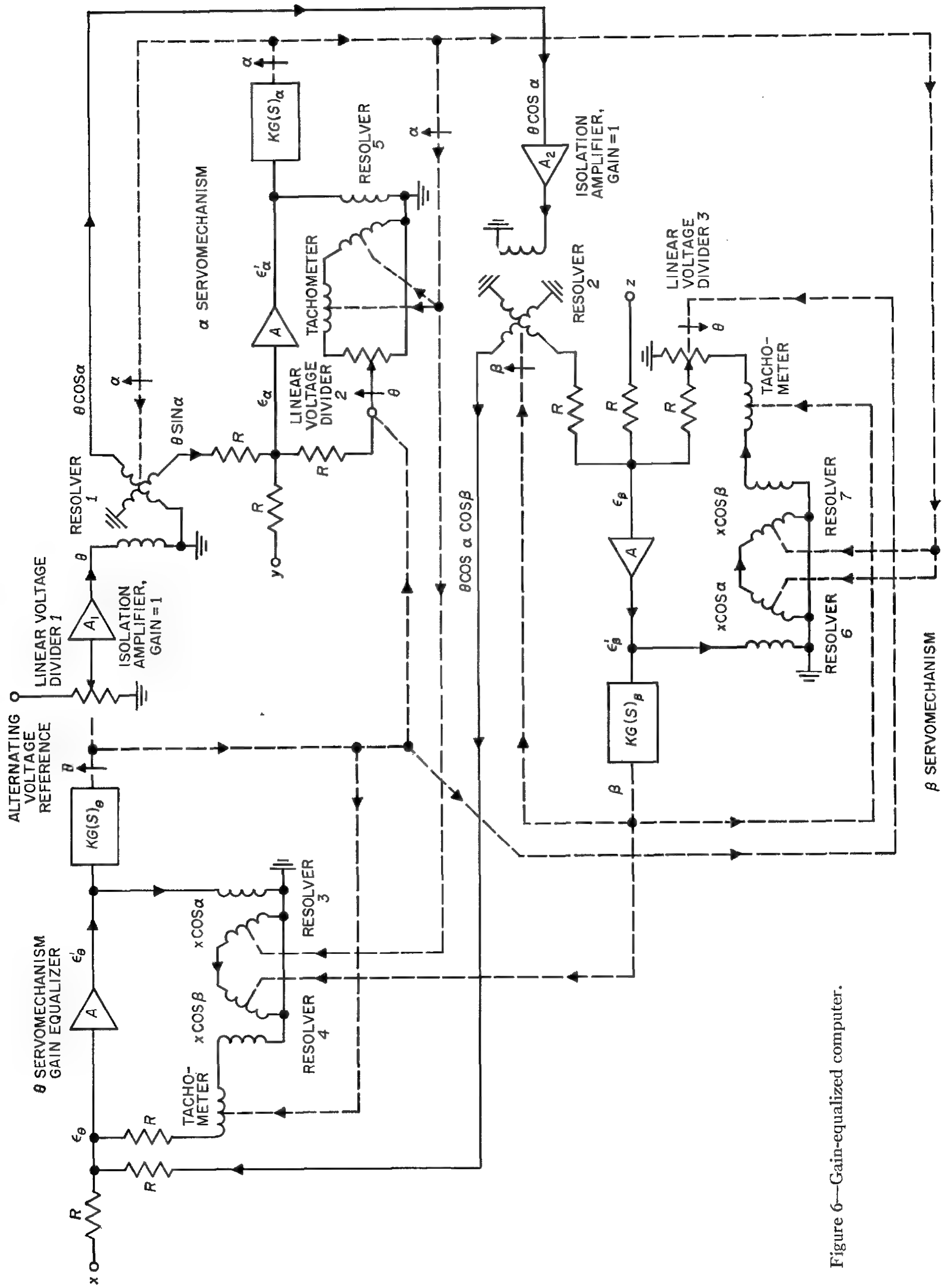


Figure 6—Gain-equalized computer.

Feedback amplifier gain-equalizing methods are often superior to other methods because:

A. The signal level at the output is generally higher.

B. Accurate elements may be used to control the gain.

C. Gain-equalization accuracy is high.

D. Method is probably simpler than other methods.

6. Multiple-Loop Nonlinear Computer

The multiple-loop nonlinear computer, Figure 6, was built for the purpose of finding the hypotenuse of a solid triangle and two of the associated angles. It contained three servomechanisms that solved three nonlinear equations. The purpose of presenting this system is to indicate the problems encountered in trying to gain equalize a multiple-loop nonlinear computer and to demonstrate the method used. Gain equalization was needed for high positional accuracy over the operating range.

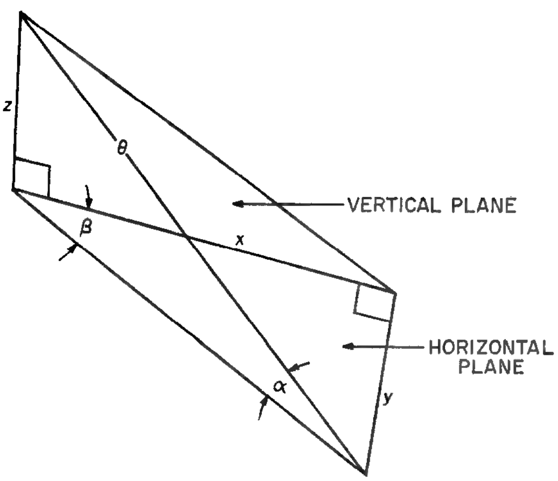


Figure 7—Vector diagram.

Referring to Figure 7, the equations solved, written to indicate the error, were

$$x - \theta \cos \alpha \cos \beta = \epsilon_\theta \tag{29}$$

$$y - \theta \sin \alpha = \epsilon_\alpha \tag{30}$$

$$z - \theta \cos \alpha \sin \beta = \epsilon_\beta. \tag{31}$$

The inputs to the equations are the quantities x, y, and z. The servomechanism outputs are θ, α, and β. It may be seen that in general, α, β, and θ all respond to any input of x, y, or z. Therefore the error sensitivity of the θ servomechanism, for example, involves α and β in addition to θ. But α and β are time functions. Consideration must be given both to the gains as determined from the foregoing equations and to the stability.

The procedure used is first to find the derivatives in (26), (27), and (28) from which a gain equivalent circuit can be made. The equivalent circuit is studied to determine what steps are necessary for gain equalization and system stability.

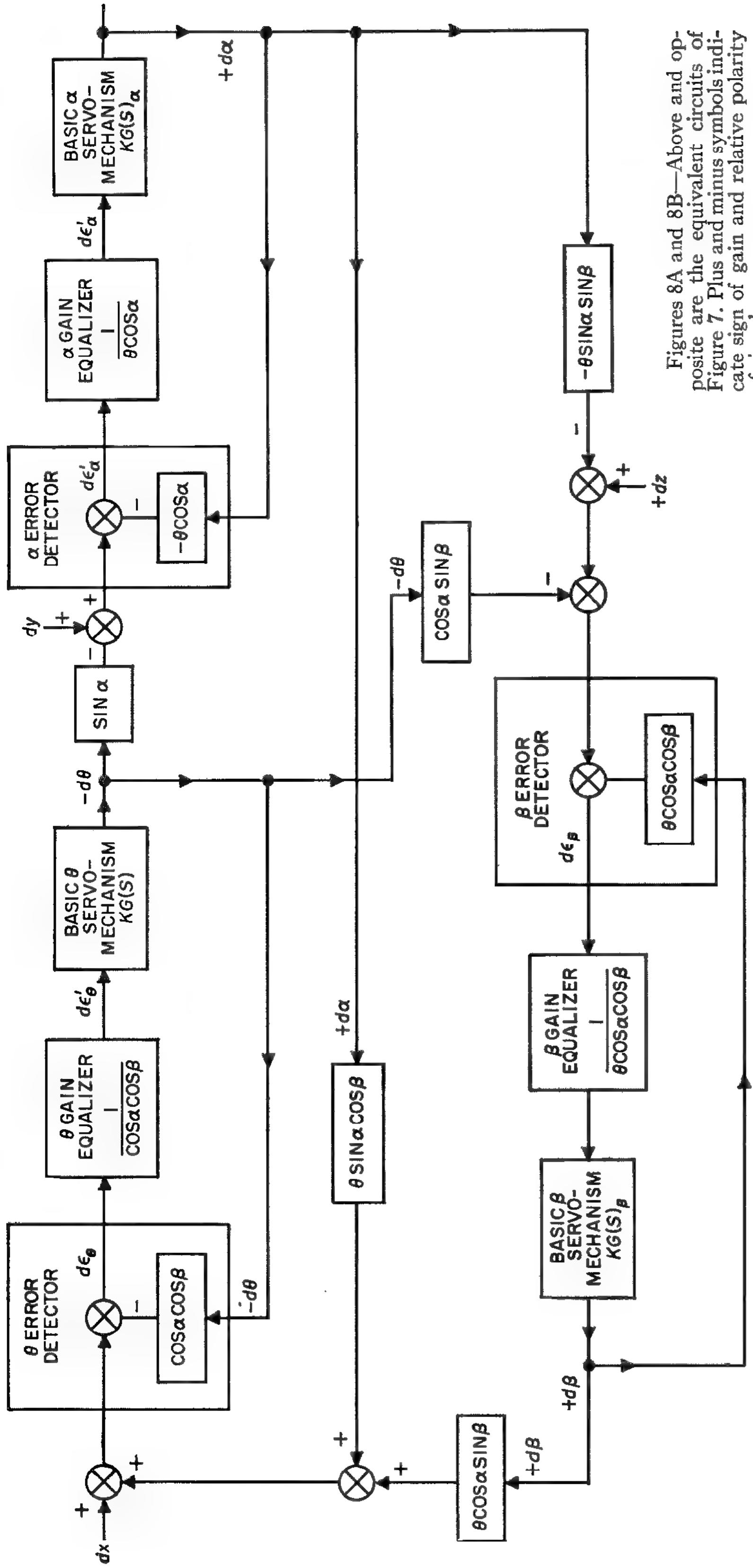
The first step in obtaining the equivalent circuit is to trace out and understand the basic operation of the computer. Considering first the primary signal path, a voltage proportional to the rotation θ is obtained from the θ servomechanism and applied to a linear voltage divider, the output of which is an alternating voltage proportional to θ. The θ voltage is applied by isolation amplifier A_1 to resolver *1*, which is positioned by the α servomechanism at angle α. The outputs are voltages $\theta \cos \alpha$ and $\theta \sin \alpha$.

The $\theta \cos \alpha$ voltage is applied to isolation amplifier A_2. The output of A_2 is applied to resolver *2*, which is positioned at angle β by the β servomechanism. The voltages appearing at the resolver output windings are $\theta \cos \alpha \cos \beta$ and $\theta \cos \alpha \sin \beta$.

Voltages proportional to x and $\theta \cos \alpha \cos \beta$ are subtracted at the input to the θ servomechanism by the two adding resistors to form the error voltage ϵ_θ. The θ servomechanism maintains the error voltage ϵ_θ at zero by controlling the θ quantity in the $\theta \cos \alpha \cos \beta$ voltage from resolver *2*.

The α servomechanism operates similarly. The input y and the feedback quantity, $\theta \sin \alpha$ from resolver *1*, are subtracted by two adding resistors to form the error voltage ϵ_α. The α servomechanism positions resolver *1* to control the magnitude of $\theta \sin \alpha$. Equation (30) is thereby solved.

The β servomechanism solves (31). The input to the β servomechanism is voltage z, and the



Figures 8A and 8B—Above and opposite are the equivalent circuits of Figure 7. Plus and minus symbols indicate sign of gain and relative polarity of signals.

A. GAIN-EQUIVALENT CIRCUIT WITH INDIVIDUAL SERVOS EQUALIZED

feedback quantity is voltage $\theta \cos \alpha \sin \beta$. These two voltages are subtracted by the adding resistors to form error voltage ϵ_β. The β servomechanism positions resolver 2 to maintain ϵ_β at zero.

Circuits for gain equalization and damping will be described after consideration of gain equalization requirements.

Taking the derivatives of (29), (30), and (31), one obtains

$$d\epsilon_\theta = dx - \cos\alpha \cos\beta \, d\theta + \theta \sin\alpha \cos\beta \, d\alpha + \theta \cos\alpha \sin\beta \, d\beta \quad (32)$$

$$d\epsilon_\alpha = dy - \sin\alpha \, d\theta - \cos\alpha \, d\alpha \quad (33)$$

$$d\epsilon_\beta = dz - \cos\alpha \sin\beta \, d\theta + \theta \sin\alpha \sin\beta \, d\alpha - \theta \cos\alpha \cos\beta \, d\beta. \quad (34)$$

Equations (32), (33), and (34), with the aid of Figure 7, may be combined as an equivalent circuit, Figure 8A. The individual servomechanisms of Figure 8A are gain equalized within themselves by assuming that the feedback paths from the other servomechanisms are opened.

Referring to the gain-equalized θ servomechanism, which has a feedback factor $\cos \alpha \cos \beta$ within the error detector, the zero-frequency output–input response is

$$\left.\begin{aligned}\frac{d\theta_o}{d\theta_i} &= \frac{\dfrac{KG(s)_\theta}{\cos\alpha\cos\beta}}{1+\dfrac{KG(s)_\theta}{\cos\alpha\cos\beta}\cos\alpha\cos\beta}\\ &= \frac{\dfrac{KG(s)_\theta}{\cos\alpha\cos\beta}}{1+KG(s)_\theta}\end{aligned}\right\} \quad (35)$$

$$= 1/(\cos\alpha\cos\beta)/w = 0.$$

The loop gain, before gain equalization, is proportional to $\cos\alpha\cos\beta$ if it is assumed that the other servomechanisms are disabled. This may be determined by inspection of (32) in which dx, $d\alpha$, and $d\beta$ would become zero. The reciprocal of this function is $1/(\cos\alpha\cos\beta)$, the required gain-equalization function.

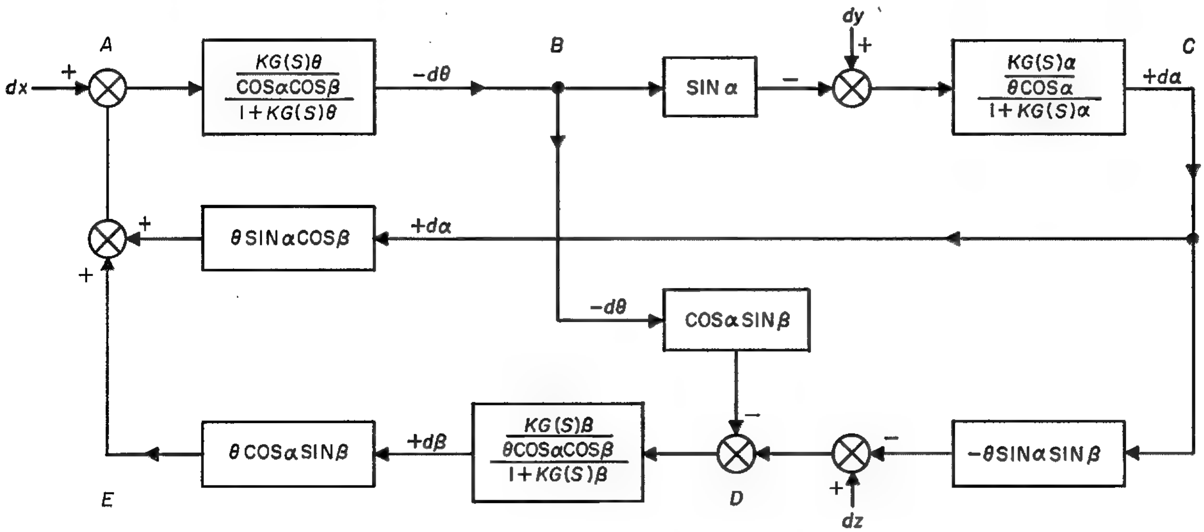
In the θ servomechanism loop, the gain equalization is provided by feedback around a

(30) and (31). The responses are

$$\frac{d\alpha_o}{d\alpha_i} = \frac{\dfrac{KG(s)_\alpha}{\theta\cos\alpha}}{1+KG(s)_\alpha} = 1/(\theta\cos\alpha)/w = 0, \quad (36)$$

$$\frac{d\beta_o}{d\beta_i} = \frac{\dfrac{KG(s)_\beta}{\theta\cos\alpha\cos\beta}}{1+KG(s)_\beta} = 1/(\theta\cos\alpha\cos\beta)/w = 0. \quad (37)$$

The gain-equalization factors, as determined from (33) and (34), are included in (36) and (37). The factors are the reciprocals of the error sensitivities, $1/(\theta\cos\alpha)$ for the α servomechanism and $1/(\theta\cos\alpha\cos\beta)$ for the β servomechanism, with the other servomechanisms disabled in each case. In the α system, the feedback factor is obtained by using a high-gain amplifier and feed-



B. FIRST SIMPLIFIED GAIN-EQUIVALENT CIRCUIT

Figure 8B.

high-gain amplifier. The feedback factor $\cos\alpha\cos\beta$ is obtained from resolvers *3* and *4*. The induction generator (tachometer), the output of which is in series with the output of resolver *4*, provides variable damping proportional to $1/(\cos\alpha\cos\beta)$. The need for this is explained later. In this instance it is intended that $KG(s)_\theta$ represent an undamped open-loop servomechanism.

The output–input responses of the α and β servomechanisms are similarly determined from

back proportional to $\theta\cos\alpha$ from resolver *5* and voltage divider *2*. Similarly in the β system, requiring a feedback factor $\theta\cos\alpha\cos\beta$, the feedback is obtained using resolvers *6* and *7* and voltage divider *3*. The outputs of the tachometers are connected in the feedback loop in a manner such that the damping gain is proportional to $1/\cos\alpha$ for the α system and $1/(\cos\alpha\cos\beta)$ for the β servomechanism. (This is true because at the point of application of the tachometric feedback, the equalizing voltage divider is in

the forward path and the equalizing resolvers are in the feedback path.)

The gain equalization and stability of the entire system may be studied by considering Figure 8B in which each circuit of Figure 8A has been combined into a single transfer function. Studying the zero-frequency response and stability of the system can be done with the aid of Figure 8B and (32), (33), and (34). Expressions for zero-frequency gain can be found that may be converted simply into time functions when frequency response is studied.

The gain of parallel paths $(BD + BCD)$ in series with path DE is

$$\begin{aligned} \text{gain } (BD + BDC)DE &= \left(\cos\alpha\sin\beta + \frac{\theta\sin^2\alpha\sin\beta}{\theta\cos\alpha}\right) \\ &\quad \times \frac{\theta\cos\alpha\sin\beta}{\theta\cos\alpha\cos\beta} \qquad (38) \\ &= \frac{\sin\beta\,(\cos^2\alpha + \sin^2\alpha)}{\cos\alpha} \times \frac{\sin\beta}{\cos\beta} \\ &= \frac{\sin^2\beta}{\cos\alpha\cos\beta}. \end{aligned}$$

The gain of path BCE is

$$\left.\begin{aligned} \text{gain } (BCE) &= \frac{\sin\alpha\,\theta\sin\alpha\cos\beta}{\theta\cos\alpha} \\ &= \frac{\sin^2\alpha\cos\beta}{\cos\alpha}. \end{aligned}\right\} \qquad (39)$$

The entire system loop gain is that of path AB in series with parallel paths $[(BD + BCD)DE + BCE]$.

$$\begin{aligned} \text{System gain} &= \left(\frac{\sin^2\beta}{\cos\alpha\cos\beta} + \frac{\sin^2\alpha\cos\beta}{\cos\alpha}\right) \\ &\quad \times \frac{1}{\cos\alpha\cos\beta} \qquad (40) \\ &= \frac{\tan^2\beta + \sin^2\alpha}{\cos^2\alpha} \\ &= \sec^2\alpha\tan^2\beta + \tan^2\alpha. \end{aligned}$$

Equation (40) indicates that the system gain is infinite for α or $\beta = \pm 90$ degrees; this is not true because the gain of each servomechanism is zero at these points due to limited equalizer gain. The system gain actually reaches some maximum as α and β tend to ± 90 degrees, but becomes zero at ± 90 degrees. However, the zero-frequency loop gain is negative for all loops; this may be seen from Figure 8B in which the sign of all quantities is indicated. For example, a negative change $(-d\theta)$ at point B results in positive quantities arriving at A to compensate the change.

The system stability may now be considered. Assume the angle β is zero as a result of the z input being zero. The system gain from (40) is

$$\text{gain} = \tan^2\alpha. \qquad (41)$$

This quantity becomes very high as α tends to ± 90 degrees because of the gain equalization. Consider a viscous-damped second-order servomechanism. The output–input ratio at the natural frequency is

$$\frac{\theta_o}{\theta_i}(j\omega_N) = \frac{1}{2c}\,\underline{/-90^\circ}. \qquad (42)$$

The servomechanisms in Figure 6 have variable damping such that for the three servomechanisms the damping factors are

$$c_\theta = c/(\cos\alpha\cos\beta) \qquad (43)$$

$$c_\alpha = c/\cos\alpha \qquad (44)$$

$$c_\beta = 1/(\cos\alpha\cos\beta). \qquad (45)$$

The loop gain at the natural frequency is of interest in (41). Since the total phase shift of the θ and α servomechanisms is -180 degrees, ω_N would be the frequency at which the system would oscillate. Considering (41), (42), (43), and (44), equation (41) may be rewritten

$$\begin{aligned} \text{gain } \theta_\alpha &= \tan^2\alpha\cdot(1/2c_\theta)\cdot(1/2c_\alpha)\,\underline{/-180^\circ} \\ &= (\sin^2\alpha)/4c^2\,\underline{/-180^\circ}. \end{aligned} \qquad (46)$$

Equation (46) shows that for $c > 0.5$ the system is stable for $\alpha = \pm 90$ degrees since the gain is less than unity.

Similarly if α is zero, the gain using variable damping is

$$\text{gain } (\theta_1\beta) = (\sin^2\beta)/4c^2\,\underline{/-180^\circ}. \qquad (47)$$

The path $ABCDE$ is closed during combined input conditions. This path has three servomechanisms in series and hence represents the most-unstable situation. Ignoring other paths,

the direct-current gain is

$$\left.\begin{aligned} \text{gain } (ABCDE) &= \frac{\sin^2 \alpha \sin^2 \beta}{\cos^2 \alpha \cos^2 \beta} \\ &= \tan^2 \alpha \tan^2 \beta. \end{aligned}\right\} \quad (48)$$

Here as before the loop gain tends to a maximum, being limited by the servomechanism gain of zero at ± 90 degrees. If the damping term c of each loop is increased to 0.707 or more, the frequency of greatest gain is zero and the minimum loop gain is

$$\text{gain } (ABCDE)_{\max} = \sin^2 \alpha \sin^2 \beta \cos \alpha. \quad (49)$$

The maximum value of (49) is 0.577 occurring for $\alpha = 55$ degrees and $\beta = 90$ degrees.

The computer as first constructed was unstable in ranges greater than α or $\beta = 45$ to 60 degrees. Instability was characterized by large, fairly low-frequency oscillations. This was due in part to velocity saturation of the θ servomechanism. Application of variable damping permitted increases of α and/or β to 70 or 75 degrees before instability would occur. The loop gain with, say, $\beta = 0$ is proportional to $\tan^2 \alpha$, which was increased by factors ranging from 2.5:1 to 14:1 by use of variable damping. Overdamping in individual servomechanisms, although not desirable, generally was not serious because α, β, were usually small quantities in the intended application of the system.

7. Symbols Used

A = gain
c = damping factor
E = reference voltage
$KG(s)$ = basic open-loop servomechanism transfer function
R = resistance
R_p = voltage-divider resistance
s = Laplace transform variable
x = input variable
y = input variable
z = input variable
α = output rotation
β = output rotation
ϵ = error produced by error detector
ϵ' = error produced by gain equalizing circuit
θ_i = input rotation
θ_o = output rotation
θ_0 = rotation as fractional portion of total resistance.

Subscripts o and i refer to output and input, respectively.

Compandor System Z6NC for Short-Haul Carrier Telephony*

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ELECTRICAL and constructional details of the *Z6NC* system are briefly described. Based on diagrams, the description gives the special features of the *Z6NC* system: compandor properties; effect of compressed, normal, and expanded speech on crosstalk; level control; and other features. In conclusion, the most-important system properties for application to cable and open wire line are given.

• • •

Six telephone calls can be transmitted on a two-wire system by the *Z6NC* compandor carrier-telephone system. Each channel has its own compandor, improving transmission quality by lessening the effects of noise and crosstalk. Net loss stability is better than that specified by recommendations of the Comité Consultatif International Téléphonique since automatic level control is incorporated in each channel. The system permits transmission of dialing or metering pulses with small time delay and low distortion. Considerable simplification in design is obtained by utilization of the compandor, which also permits simultaneous carrier transmission on both side and phantom circuits. In addition to the usual application to cables, the system can be applied to open wire lines since automatic level control is provided in both channel and group equipments and in the repeaters.

1. Introduction

The *Z6NC* carrier system may be regarded as a further development of the *Z6NT* system introduced in 1952,[1] which also transmits six carrier-frequency calls on a two-wire line in deloaded or nonloaded symmetrical-core cable and is designed primarily for the lower network, the arrangement of lines between the end office and the primary center. However, the newer system is further suitable for application to open wire lines or aerial cables.

There are three points in which the *Z6NC* system is superior to the older system. These are:

A. Improved signal-to-noise ratio.

B. Incorporates signal transmission circuits.

C. Greater stability of net loss.

Development of the *Z6NC* was carried out with close adherence to specifications for short-haul systems issued by the Deutsche Bundespost.

2. Circuit Applications

The introduction of a compandor[2] permits telephone operation under very-unfavorable noise and crosstalk conditions. Moreover, the compandor with its improvement of signal-to-noise ratio by 2.6 nepers* enables simultaneous use of the side and phantom circuits of star-twisted and multiple-twin cables enabling transmission of 18 calls on a quad. Even in marginal cases, simultaneous utilization of side and phantom circuits can be secured by inexpensive phase shifters inserted in the carrier generator at the terminal.[3] Crosstalk compensation in lines with normal crosstalk, even in lines of older types, is unnecessary.

The system carries signals with satisfactory speech immunity, permitting transmission of dialing and metering pulses during conversation. Due to level control, the signal distortion is at a minimum.

The automatic level regulation for each channel maintains the specified toll-circuit net loss despite attenuation variations of ± 1 neper.

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* Reprinted from *FTZ*, volume 8, pages 502–511; September, 1955.

[1] L. Christiansen, "Nahverkehrs-Sechskanal-Tragerfrequenz-System *Z6NT*," *SEG-Nachrichten*, volume 2, pages 4–12; 1953.

[2] M. Jänke, E. Prenzel, and W. Speer, "Dynamikpresser und -dehner für Fernsprechverbindungen," *FTZ*, volume 6, pages 459–469; October, 1953.

* 1 neper = 8.686 decibels.

[3] W. Hofmann, "Nebensprechausgleich bei Zweiseitenbandsystemen," *FTZ*, volume 8, pages 555–558; October, 1955.

In addition to the level regulation of each channel, level regulation for the over-all group can be switched in. This permits utilization of the system over open wire lines with or without cable lead-ins or over aerial cables. In these cases, attenuation variations of up to approximately 4 nepers at the highest frequency can be compensated. The compandor also permits utilization of open wire lines with a high amount of interfering noise (for instance, interfering radio transmitters). It may be stressed here that no special rearrangement within the system is necessary for application to open wire lines; it is ready for operation after changing a few connections.

Automatic level regulation or, if necessary, group frogging can be applied in *Z6NC* repeaters that are suitable both for cable and open wire operation.

Compatibility with *Z6N*, *Z12N*, and *Z12K* systems operated on adjacent circuits, for instance in the same cable, is easily established by shifting the lower transmission band by 2 kilocycles per second. The carrier frequencies of the *Z6NC* system then coincide with the zero frequencies of the other systems.

The equipment is produced in the new mechanical design of the Deutsche Bundespost with plug-in panels.

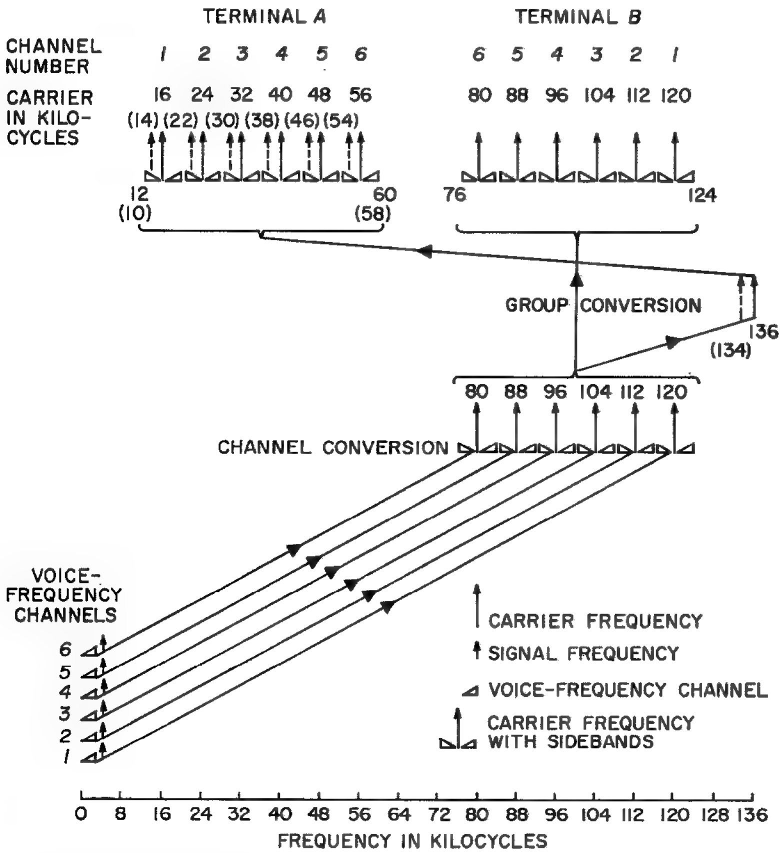


Figure 1—Modulation scheme. Substitution of the 136-kilocycle group conversion carrier by 134 kilocycles drops the lower hand by 2 kilocycles for compatability with other systems.

3. *Electrical Details*

3.1 Terminal Equipment

3.1.1 Modulation Scheme

Figure 1 shows the modulation scheme of the transmitting terminals. Voice-frequency channels *1* through *6* and the accompanying signal frequencies are modulated with carriers of 80 through 120 kilocycles and thus appear in the desired frequency band at terminal *B*. The group carrier frequency of 136 or 134 kilocycles translates the high band into the desired low-frequency band for terminal *A*.

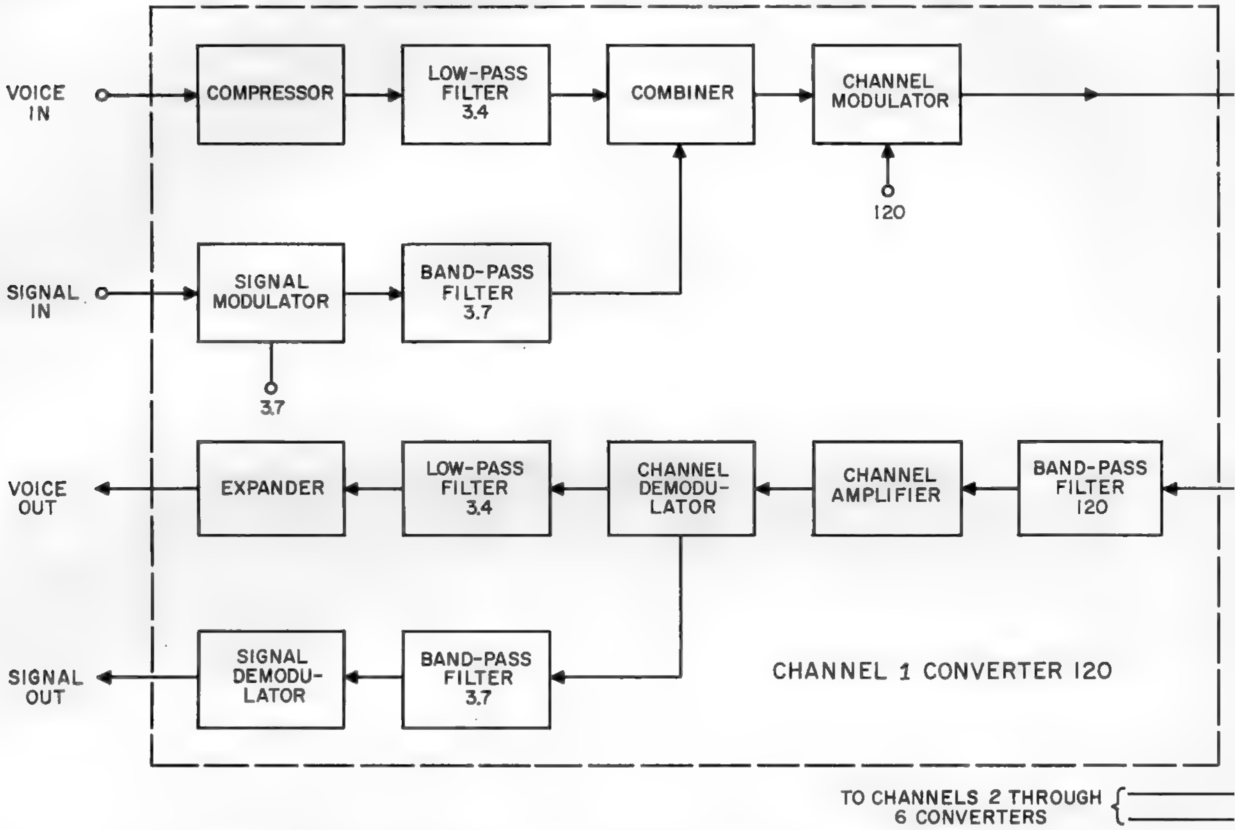
Channel demodulation as well as channel modulation is done in the 80-to-120-kilocycle band. At the receiver, the individual carrier-frequency channels are separated by simple band-pass filters, amplified, and demodulated by full-wave rectification.

3.1.2 Transmitting and Receiving Paths

The transmitting and receiving paths are shown in Figure 2. Only the essential points of these circuits will be pointed out here. A dynamic compressor is located at the channel input and a dynamic expander at the output. These form the previously mentioned compandor. The signal and speech circuits are combined before channel modulation. The filter usually found in the output

circuit of the channel modulator has been omitted owing to sufficient decoupling and to the action of the compandor. The signal modulator passes the signal frequency of 3.7 kilocycles when the signal input conductor is grounded. The signal spectrum generated by this keying is amplifier. (If necessary, the level of the transmitted high band can be increased to 1.5 ± 0.3 nepers.)

The receiving carrier level must be between + 0.5 and − 8.6 nepers at 120 kilocycles. In the receiving path, the incoming frequency band



limited by the band-pass filter. The channel modulator output contains the carrier frequency plus all sidebands produced by the voice and signal frequencies. The depth of modulation of the carrier is about 35 to 45 percent for both the signal and speech currents when zero level (1 milliwatt in 600 ohms) is applied to the channel input.

The six channels, decoupled by attenuators, are merged before group translation. The low-pass filter at the converter input attenuates the higher-order modulation products to a sufficiently low value. The group modulator translates the high band into the low band. The individual carriers at the output of the directional filter have a power level of 0.5 neper, variable by ± 0.3 neper in steps of 0.1 neper in the transmitting

(16 to 124 kilocycles in the case shown in Figure 2) is applied to an equalizer, pad, and receiving preamplifier so that all carrier levels are of equal magnitude. Due to the double-sideband transmission, the requirement of accurate equalization is not too stringent here. The main receiving amplifier next in line has a gain independent of frequency and a low internal resistance; the incoming carrier levels can be measured at its output, where a jack is provided. The individual channels are separated by band-pass filters; the only constructional units in which the channels differ. The channel amplifier is automatically gain-regulated, this regulation responding to the magnitude of the carrier level. The carrier is demodulated next. The direct current obtained from the channel demodulator

has a magnitude dependant on the carrier level; it operates a control relay indicating absence of carrier or any other discontinuity in the transmission path. The (compressed) voice-frequency band obtained by demodulation is applied via a low-pass filter to the expander, where the dynamic range is enlarged. The output amplitude of the speech channel can be varied by a control in the expander. The range of this adjustment is $\pm$ 0.35 neper, the adjustment being stepless. If a signal is sent from the far-end terminal station, the incoming signal current passes successively through the signal band-pass filter, a rectifier demodulator circuit, and is finally applied to a polarized relay; the contact of the latter grounds the output signal conductor.

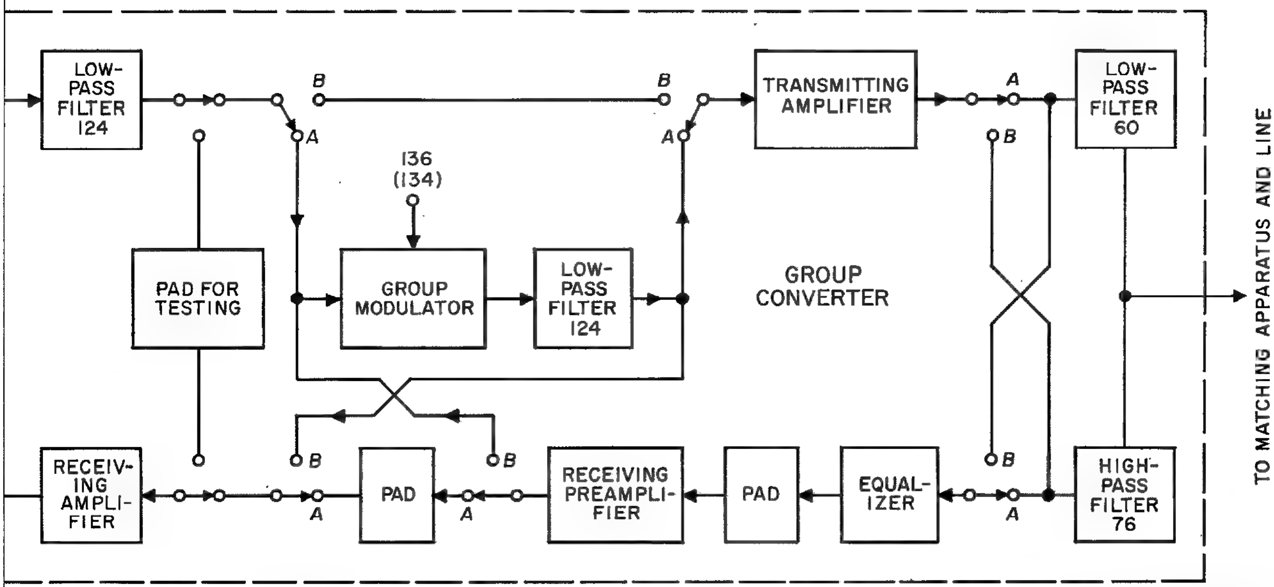


Figure 2—Block diagram of *Z6NC* terminal equipment. The unlabeled numbers indicate frequencies in kilocycles. Connections *A* are made when the low band is transmitted and the high band received, and *B* are made when the high band is transmitted and the low band received.

Measurements of the channel modulator can be taken by connecting the test pad in the group converter.

For the transmission of voice-frequency carrier telegraph, the compandor is rendered ineffective by simple switching means.

3.1.3 Carrier Generator

The fundamental frequency, produced by an 8-kilocycle crystal oscillator, is converted into a pulse train supplying, through band-pass filters, the carrier frequencies of 80 to 120 kilocycles required for channel modulation. The oscillator generating 3.7 kilocycles for a signal modulator has no crystal. The group carrier frequency of 136 (or 134) kilocycles is again obtained from a crystal oscillator, the change from 136 to 134 kilocycles being accomplished by exchanging crystals. The carrier-supply capacity is so dimensioned that four terminal stations making up one bay can be furnished with the carrier voltages required. The group carrier generator can also supply energy for 8 repeaters that can be accommodated in a bay of the same size.

3.1.4 Power Supply and Tube Complement

The power for a complete bay with four terminal stations is supplied by the mains or partly by batteries. The power supply unit is therefore included in the bay; its power consumption amounts to approximately 670 volt-amperes. The tube complement for four terminals including carrier power supply consists of 97 tubes (*C3m* pentodes with a guaranteed life of 10 000 hours of operation).

3.2 Repeater

The *Z6NC* repeater is provided for amplification of the 12-to-60- (or 10-to-58-) and 76-to-

124-kilocycle frequency bands, transmitted in two-wire operation. To avoid feedback of cross-talk through other circuits, the incoming transmission bands are interchanged at the repeater output. The frequency-dependent attenuation of short-haul cables up to values of 8.3 nepers at 124 kilocycles (corresponding to 33 kilometers, or 21 miles of 1.4-millimeter copper conductor) can be compensated by the repeater. The equipment can be operated as a two-wire repeater without group frogging also, and can operate as a four-wire repeater with or without frogging.

In Figure 3, the incoming band is equalized by an equalizer, pad, and preamplifier. Subsequently, the band is translated. The band thus obtained is amplified in the main amplifier with a large amount of feedback and appears across the toll line via the low- or high-pass part of the output directional filter and the toll-line transformer.

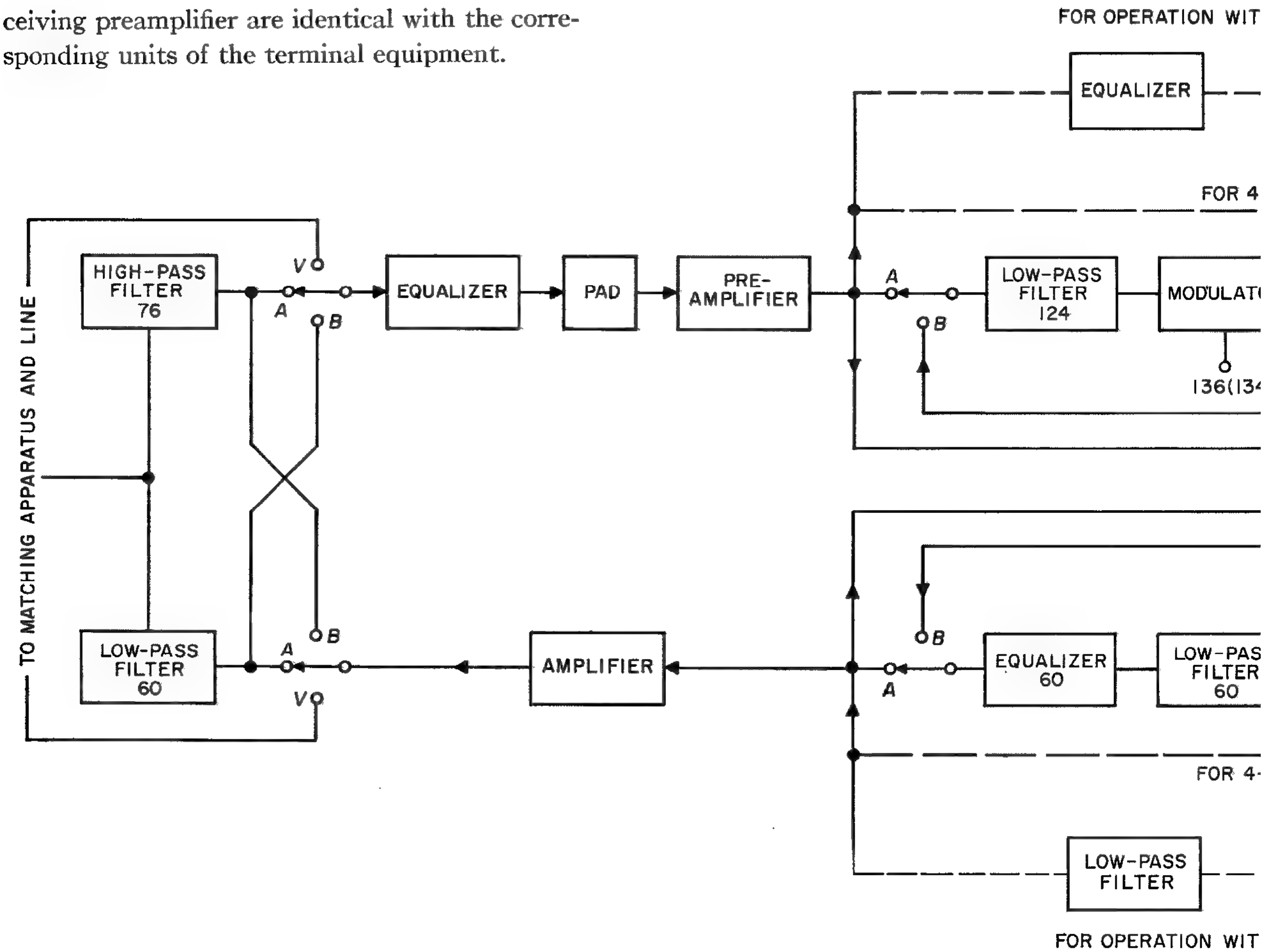
The directional filter, equalizer, pad, and receiving preamplifier are identical with the corresponding units of the terminal equipment.

4. Mechanical Construction

The *Z6NC* system is mounted in exchangeable plug-in panels in bays with lock and key according to the mechanical design of the Deutsche Bundespost.

The standard bay generally takes four complete terminal equipments each comprising three channel-converter panels (for two channels each) and a group-converter panel. The carrier supply consists of one panel and a unit accomodated on the operator's board. This, as well as the panel with heater-power and signaling-current supply and the panel with the plate-current ($B+$) supply, are provided centrally and are so dimensioned that a fully equipped bay with four terminals can be supplied with power.

Figure 4 shows an opened cabinet with 24 channels. The bay is so wired that terminals alone or repeaters alone or both arbitrarily mixed



can be accommodated. When fully equipped, the bay will take 8 repeaters.

Constructional details of the channel modulator panel are shown in Figure 5. The photograph shows the complete panel and several subunits.

The control jack strips of the panels comprise level and other measuring jacks and circuit-breaker plugs. Signals indicate any breakdown of operational voltages, fuses, or tubes.

The bay is about 2600 millimeters (102 inches) high, 600 millimeters (24 inches) wide, and 222 millimeters (9 inches) deep. The weight of a fully equipped bay is about 330 kilograms (726 pounds). If only one or two *Z6NC* terminals or repeaters are required for the completion of a line, they can be mounted in a halfbay 1500 millimeters (59 inches) high, with width and depth as for the normal bay. The weight of the halfbay is about 180 kilograms (396 pounds).

5. *Technical Innovations*

5.1 Compandor

5.1.1 General

The introduction of compandors in various international systems (for instance, the *N1*, *45A*, *ON*, and *O* systems) aims at improvement of transmission properties without substantially increasing the cost.

Various publications[2,4,5] contain detailed descriptions of the effect of the compandor, its properties, and its performance when handling speech or sinusoidal signals.

In the following, an attempt is made to describe briefly the most important properties of such equipment.

The two components of the compandor; that is, the compressor and the expander; are electrically so designed that no difference in the speech

[4] R. S. Caruthers, "N-1 Carrier System," *Bell System Technical Journal*, volume 30, pages 5–32; January, 1951.

[5] G. Hässler, "Sprachübertragung mit Dynamikkompression," *FTZ*, volume 12, pages 659–664; December, 1954.

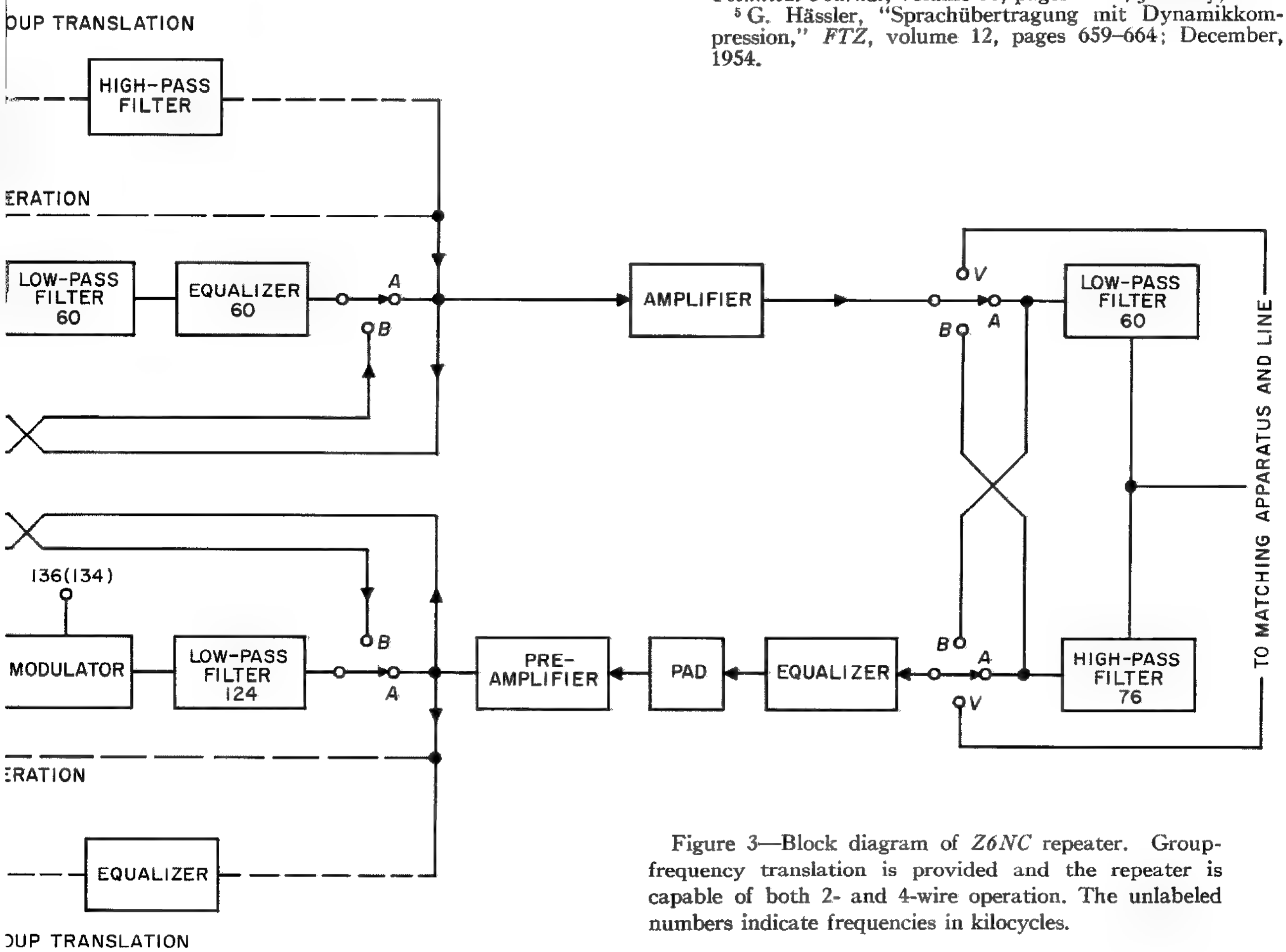


Figure 3—Block diagram of *Z6NC* repeater. Group-frequency translation is provided and the repeater is capable of both 2- and 4-wire operation. The unlabeled numbers indicate frequencies in kilocycles.

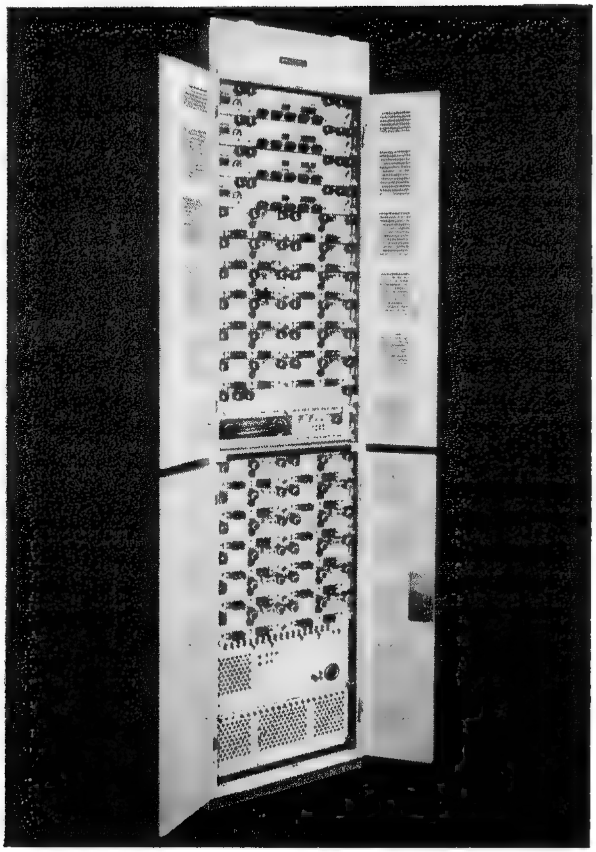


Figure 4—Front view of *Z6NC* bay equipped with 4-terminal equipment.

signals between the system input and output can subjectively be found.

Even when 10 compandors are connected in series, the loss of speech quality is not deleterious. The intelligibility at the output of a system depends both on the noise component superimposed on syllabic speech and on the noise level during the intersyllabic periods. Both of these conditions are affected advantageously by the compandor.

During conversation, the compressor emphasizes low-volume syllables. Hence, transmission on the toll line is done at a higher average level; the low-volume syllables ride above the noise level in which they would be drowned in a similar system without a compandor. This alone leads to speech of better quality.

Between speech intervals, the expander causes considerable suppression of interfering noise (standard value of about 2.6 nepers); speech following a quiet interval thus becomes more intelligible due to a peculiarity of the human ear, which adapts itself to higher sensitivity during a quiet interval.

With respect to equipment design, the following advantages are obtained by the use of the compandor: Lower requirements on linearity of amplifiers and modulators. Reduced requirements with respect to the blocking properties of filters in speech and signal circuits. Elimination of channel transmitting filters. More compact construction because of the reduced danger of mutual interference between components.

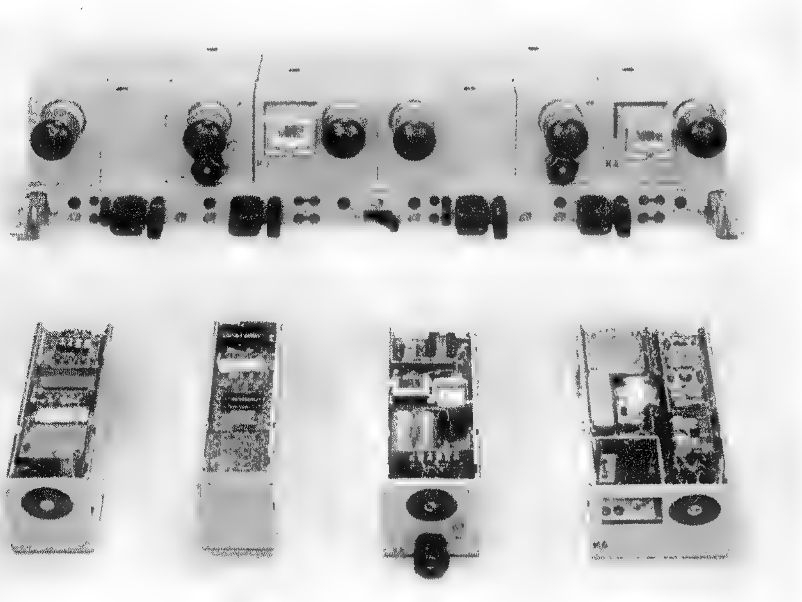


Figure 5—Channel converter panel with subassemblies.

5.1.2 Electrical Characteristics of Compandor

The block diagram of Figure 6 shows the action of the compressor and expander with the characteristics below. The expander characteristic is used directly to derive the value of the expander gain (Figure 7), which depends on the

noise level. Due to the germanium diodes employed in compressor and expander, the characteristics are to a small degree dependent on temperature. A large portion of the development work was devoted to this problem. No noticeable change of the characteristics described is caused by ambient temperatures of + 10 to + 35 degrees centigrade.

The stability of short-haul systems depends on the maintenance of a defined net loss. Since the net loss in a compandor system depends somewhat on the input signal amplitude, an increase of net loss will be observed at the edges of the channel transmission band (frequency dependence) and with decreasing signal amplitudes (amplitude dependence). Figure 8 shows typical amplitude dependence for a compressor–expander system. These properties result, for the case of nonmodulated carriers and at the rated noise level, in an important improvement of the singing margin as shown in Figure 9.

A typical median characteristic of the frequency-dependent net loss for 12 channels in both transmission directions, measured with the nominal level, is plotted in Figure 10. Actual conditions are much-more favorable because the effective control current for compressors and expanders is determined by the total energy

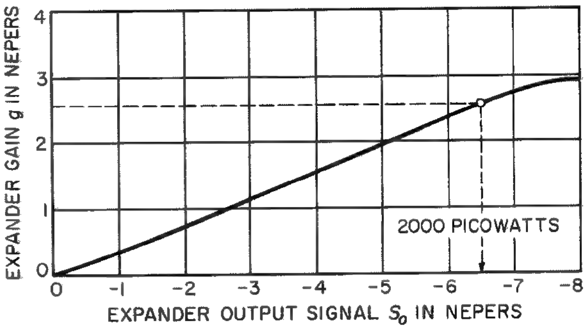


Figure 7—Expander gain characteristic. The dashed lines show the limit set by noise.

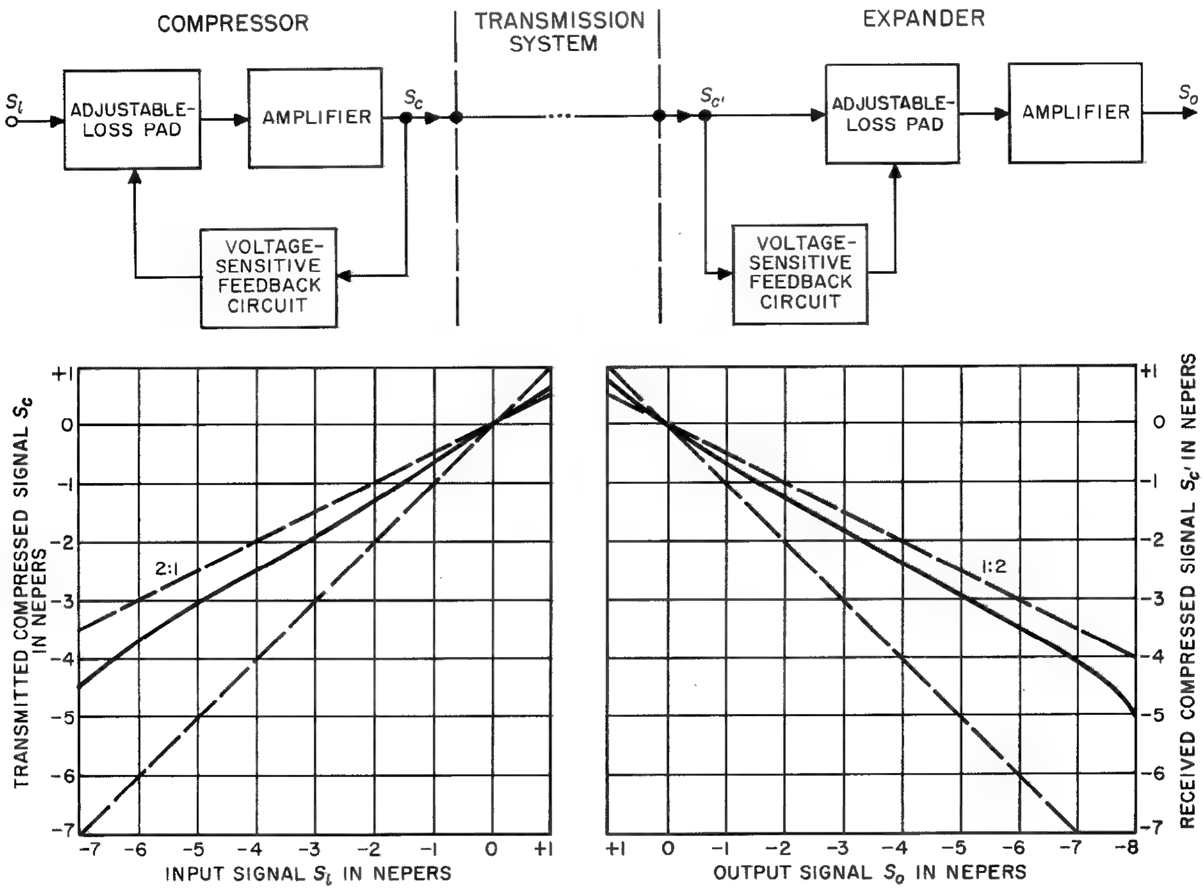


Figure 6—The compandor circuit and its characteristics with respect to relative zero level.

concentration of human speech in the range from 500 to 1000 cycles. Therefore, the net-loss variations to be expected are within 20 percent of the tolerance specification of the Comité Consultatif International Téléphonique.

5.1.3 *Effect of Compandor on Perturbances*

As long as the line noise equally affects both sidebands of the transmitted carrier, an expander gain as in Figure 7 becomes effective during the speech intervals. Since the voltages of the two sidebands are added, while the powers of the noises are added, the sideband-level-to-noise ratio on the toll line may be smaller by (0.35 neper + expander gain g) than the desired signal-to-noise ratio in the output.

For selective disturbers affecting one sideband only, the signal-to-noise ratio in the output is larger by (0.7 neper + g) than the ratio on the line.

Figure 11 is a comparison of the *Z6NC*, *Z12N*, *Z6N*, *Z12K*, and *Z6NT* systems with respect to signal-to-noise ratio in the output, assuming that line lengths are equal and noise level is constant. The comparison shows in each case a gain in signal-to-noise ratio for the *Z6NC* system. In addition, the transmission level of the high band (the band of interest here) of the *Z6NC* system can be increased to about 1.5 nepers for operation between two terminals; an additional reserve is thus available for lines greatly affected by noise.

When the crosstalk of adjacent line circuits is considered, a difference should be made between compressed and normal speech, and between speech expanded and not expanded in the receiving system. Typical cases can thus be derived as described in the legend of Figure 12.

Figure 12 is a comparison of crosstalk between

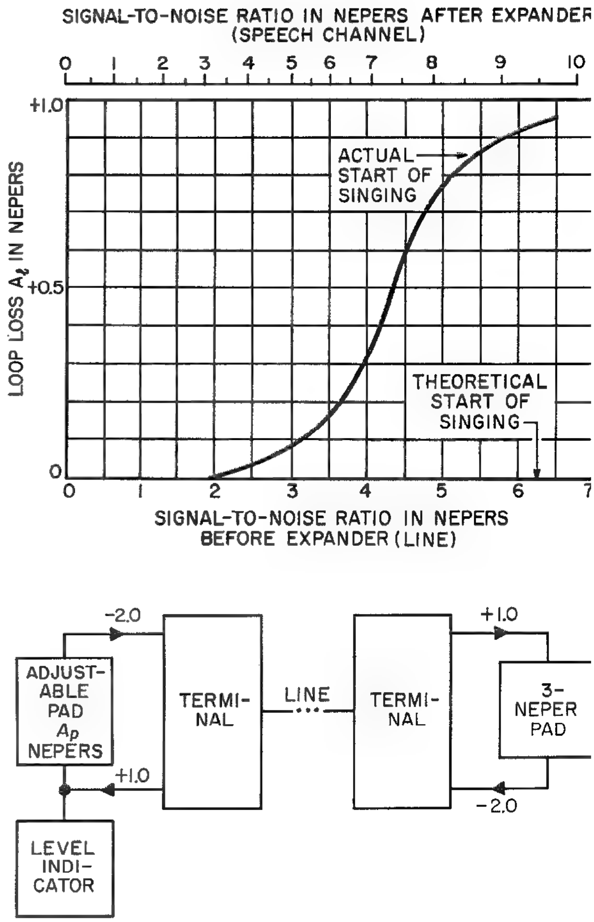


Figure 9—Loop loss as a function of noise level. Input and output levels in nepers are shown on the block diagram: Loop loss, $A_l = 3 - A_p$ nepers.

the *Z6NC* and some other well-known systems. Better intelligibility of compressed speech involves an increase in crosstalk attenuation A_c. For this effect (which depends on the degree of compression) various values (0.57 to 0.8 neper) are shown[6,7] in the literature. Subjective ob-

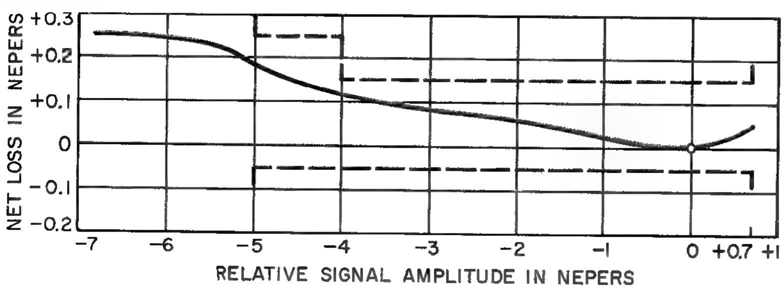


Figure 8—Compandor amplitude dependence. Dashed lines give the tolerance.

[6] F. S. Boxall and R. S. Caruthers, "Miniature Compandor for General Use in Wire and Radio Communication Systems," *Transactions of the American Institute of Electrical Engineers*, volume 72, part 1, pages 804–811; January, 1954.

[7] A. J. Aikens and C. S. Thaeler, "Control of Noise and Crosstalk on N1 Carrier Systems," *Transactions of the American Institute of Electrical Engineers*, volume 72, part 1, pages 605–610; January, 1954.

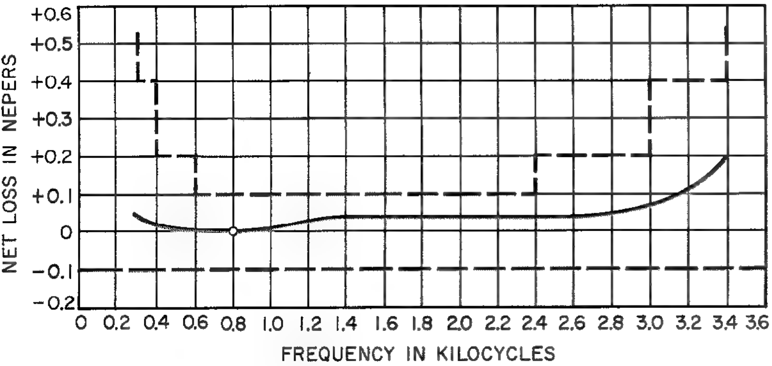


Figure 10—Net loss as a function of voice modulation frequency. The upper dashed line is the 40-percent tolerance of the Comité Consultatif International Téléphonique.

servation of German talkers has shown that A_c with compressed speech must be higher by about 0.6 neper to create the same crosstalk sensation as with uncompressed (normal) speech; this value was taken as the basis for the representations in Figure 12.

5.2 Signal Transmission

As is the case in many of the newer[4] carrier-frequency systems, the *Z6NC* system transmits

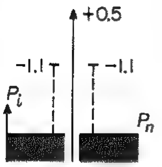
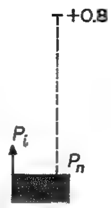
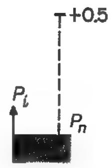
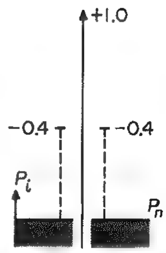
	Z6NC $m=40$ Percent	*Z12N* (*Z6N*)	*Z12K*	*Z6NT* $m=50$ Percent
Signal and Noise Level on Line in 3100-Cycle Bandwidth				
A_c Signal-to-Noise Ratio for Background Noise	$A_c=(-1.1+0.7)$ $-(A+0.6)$ $-(P_n+0.35)+g$ $=(-P_n-A)$ $-1.35+g$ $=(-P_n-A)+1.25$	$A_c=0.8-A-P_n$ $=(-P_n-A)+0.8$	$A_c=+0.5-A-P_n$ $=(-P_n-A)+0.5$	$A_c=(-0.4+0.7)$ $-(A+0.6)$ $-(P_n+0.35)$ $=(-P_n-A)-0.65$
	$A_c=K_1$	$A_c=K_1-0.45$	$A_c=K_1-0.75$	$A_c=K_1-1.9$
A_i Signal-to-Noise Ratio for Interfering Signals	$A_i=(-1.1+0.7)$ $-(A+0.6)-P_i+g$ $=(-P_i-A)$ $-1.0+g$ $=(-P_i-A)+1.6$	$A_i=+0.8-A-P_i$ $=(-P_i-A)+0.8$	$A_i=+0.5-A-P_i$ $=(-P_i-A)+0.5$	$A_i=(-0.4+0.7)$ $-(A+0.6)-P_i$ $=(-P_i-A)-0.3$
	$A_i=K_2$	$A_i=K_2-0.8$	$A_i=K_2-1.1$	$A_i=K_2-1.9$

Figure 11—Effective ratio A_c of signal to line noise and to interfering signals for the most unfavorable channels of 4 carrier systems operated over equal line lengths. For equal signal level and equal noise level at the systems outputs. m = percentage of modulation. P_i is an interfering signal; P_n is the background noise level. For *Z12N*, *Z6N*, *Z12K*, transmission loss = A (108 kilocycles). For *Z6NC* and *Z6NT*, transmission loss = $A + 0.6$ (120 kilocycles) all figures are in nepers. Expander gain $g=2.6$ nepers.

Figures 12A and 12B—Above and on the facing page are shown a crosstalk-ratio comparison between *Z6NC* and other systems. Crosstalk attenuation A_c figures in the table are given in nepers. The relative zero level for each system is S_0 In the sketch of compander operation at the left, normal speech in range *a* is of normal intelligibility; in range *b*, normal speech gives poorer-than-normal intelligibility. Compressed speech in range *a* gives better-than-normal intelligibility and gives normal intelligibility in range *b*.

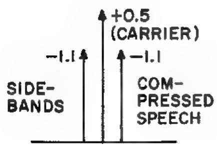
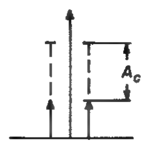
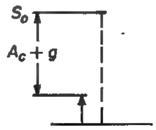
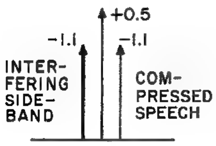
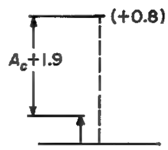
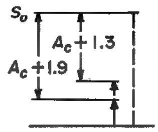
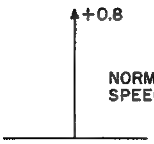
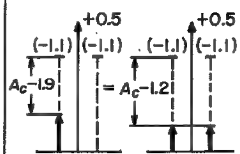
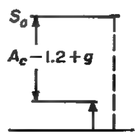
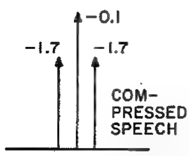
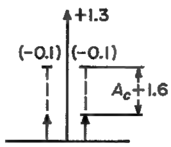
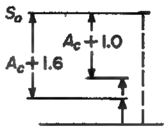
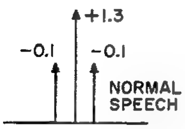
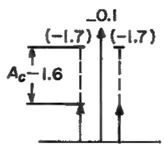
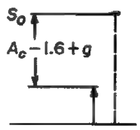
	CARRIER AND SIDE-BAND LEVEL OF TRANSMITTING SYSTEM	CARRIER AND SIDE-BAND LEVEL OF RECEIVING SYSTEM CARRIER-FREQUENCY SIDE	SIGNAL AND NOISE LEVELS OF RECEIVING SYSTEM OUTPUT	REMARKS	
A	*Z6NC*(*m*=40 PERCENT) 	*Z6NC* 	*Z6NC* 		In range *a*, Crosstalk audible as compressed speech In range *b*, Crosstalk audible as normal speech
B	*Z6NC*(*m*=40 PERCENT) 	*Z12N* 	*Z12N* 	1. Due to the better intelligibility of compressed speech, A_c decreases subjectively about 0.6 neper 2. Same ratios apply to *Z6N* 3. *Z12K* with +0.5 neper carrier and *Z6NC* with reduced carrier (+0.2 neper) give same values	
C	*Z12N* 	*Z6NC* 	*Z6NC* 	1. With expander gain of 2.6 nepers, effective A_c rises 1.4 nepers (compare S_0 for case *B*) 2. Same ratios apply to *Z6N* 3. Also for *Z12K* with +0.5 neper carrier and for *Z6NC* with −0.2 neper carrier	In range *a*, Crosstalk audible as normal speech In range *b*, Crosstalk audible as expanded speech
D	*Z6NC*(*m*=40 PERCENT) 	*Z6NT*(*m*=50 PERCENT) 	*Z6NT* 	1. Due to better intelligibility (0.6 neper), effective A_c rises 1.0 neper 2. *Z6NT* with higher carrier (1.3 nepers) and *Z6NC* with lower carrier (−0.1 neper)	
E	*Z6NT*(*m*=50 PERCENT) 	*Z6NC* 	*Z6NC* 	1. *Z6NT* with higher carrier (1.3 nepers) and *Z6NC* with lower carrier (−0.1 neper)	In range *a*, Crosstalk audible as normal speech In range *b*, Crosstalk audible as expanded speech

Figure 12A.

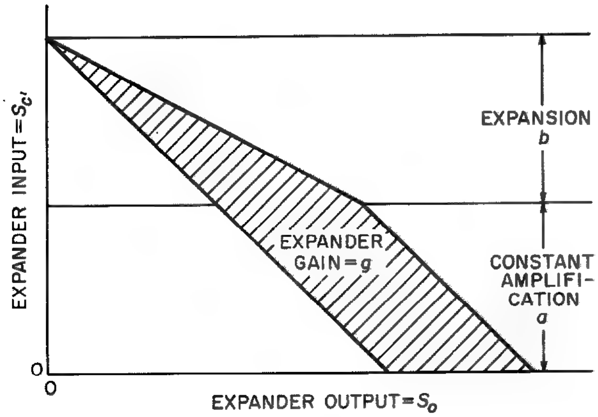


Figure 12B.

dialing signals and metering pulses during conversation through a special path in each speech channel. In signal transmission of this type, the delay time of the signals should be very short and signal distortion must be kept within certain limits. Loud speech or disturbances such as whistles must not generate undesired signal pulses. Noise in the speech channels during signal transmission must be below a definite limit.

In the system under discussion, the frequency of 3.7 kilocycles has been selected as the best compromise for out-of-band signal transmission. The requirements of negligible signal noise in speech channels, shortest signal delay times, and least signal distortion contradict each other; any rounding-off of the transmitted signal pulses for noise considerations will lengthen the delay time. The noise-suppressing effect of the expander requires only limited rounding of the pulses; this results in very-short delay times (less than 12 milliseconds). The driving current in the signaling relay is to all practical purposes independent of attenuation variations on the toll line, these variations being reduced by a factor of 12 by the channel amplifier. For instance, when the attenuation of the toll line varies by ± 1 neper, the signal distortion remains within ± 1 millisecond with a signal pulse frequency of 50/50 or 20/20 milliseconds on/off ratio.

Contact chatter of the signal relay is prevented by a special circuit causing an increase of relay current at the proper instant.

Spurious pulses caused by loud speech or whistling can be avoided with proper attenuation characteristics of the speech-channel low-pass and the signal-receiver band-pass filters. In the case of the *Z6NC* system, the specifications for these filters could be relaxed because the speech levels are reduced by a ratio of about 2 by the compressor. The curve for speech immunity (Figure 13) shows that in the most-unfavorable case a level of 1.9 nepers above relative zero level will cause a spurious signal pulse, but only in a very-narrow range of 3450 to 3550 cycles.

Dialing noise in adjacent channels with signal pulses of 50/50- or 20/20-milliseconds duration is below 0.5 millivolt measured at the system output; this is equivalent to a signal-to-noise ratio of 8.3 nepers. When the signal-modulator input is grounded, the continuous tone audible in the related channel is at least 7 nepers down; measured by a psophometer, it is 8.2 nepers down.

5.3 Level Regulation

5.3.1 Gain-Controlled Channel Amplifier

The gain-controlled amplifier of the channel balances attenuation variations in carrier-frequency transmission by regulating the carrier level. Figure 14 shows the basic circuit of the amplifier. The feedback to the amplifier is varied by the direct plate current I_p flowing through

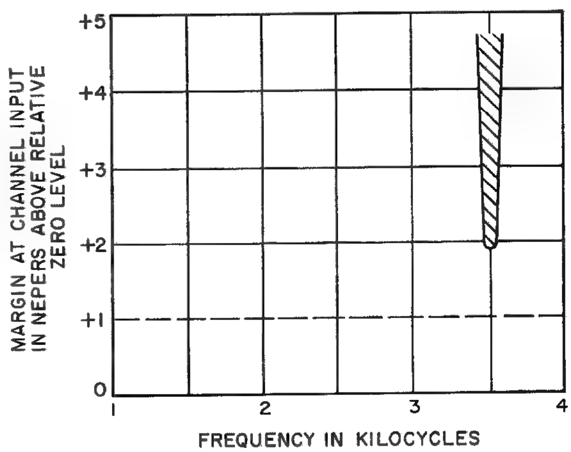


Figure 13—Speech immunity of signaling path for a channel. Shaded area is the operating region of the signal receiver; the dashed line is the specified limit. Speech immunity is the overload level applied to a speech channel at an arbitrary frequency that just fails to generate a spurious signal pulse.

a thermistor in a bridge connection. A regenerative sample of the signal in the amplifier output circuit is applied across the bridge and, depending on the bridge state of balance, a portion appears at the amplifier input in series

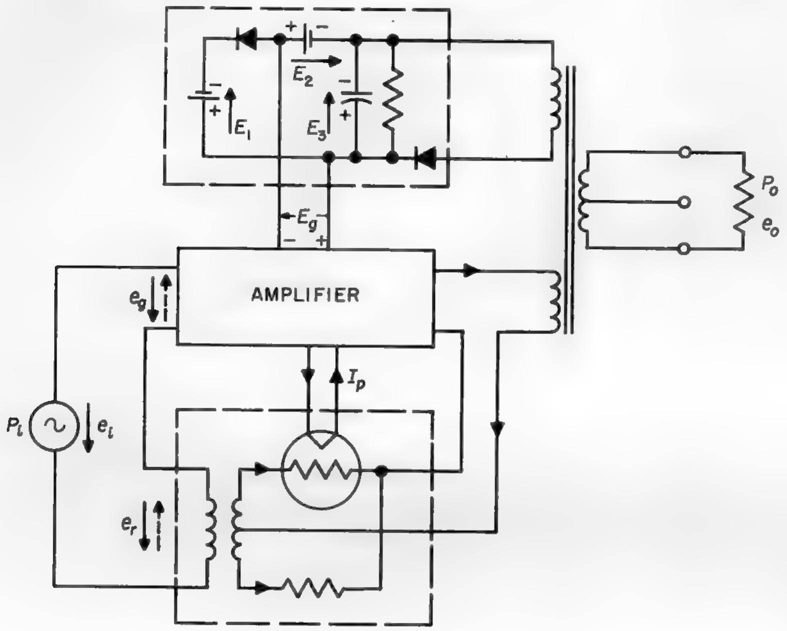


Figure 14—Variable-gain channel amplifier. Before start of control, $E_g = -E_1$ and control starts when $E_1 + E_2 = E_3$.

with input signal, e_i, as e_g. When the input level is small, the balance of the bridge is such that a regenerative voltage e_r appears at the amplifier input, the amplifier grid bias E_g being equal to E_1 and the plate current at maximum. When, with increasing input and consequently higher output voltages, E_3 reaches and exceeds the sum of E_1 and E_2, regulation begins. Grid bias E_g, when further increased, causes the plate direct current to drop and, hence, the thermistor resistance to increase. This increasing resistance changes feedback voltage e_r in the sense of increasing negative feedback. The regulation characteristic of an amplifier of this type and the variation of plate direct current are shown in Figure 15.

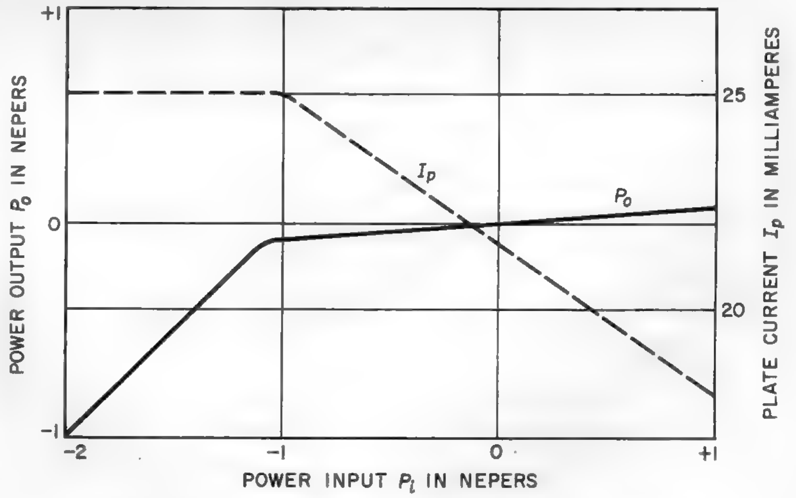


Figure 15—Characteristics of variable-gain channel amplifier.

5.3.2 Gain-Controlled Group Amplifier

The channel regulation will balance all attenuation variations occurring when the system is operated over cables. Additional variations that may be expected, for instance, on open wire lines, are compensated by additional regulation in the group amplifier. A thermistor is again used[4] as shown in Figure 16. When thermistor A is suitably preheated, a temperature-change-compensated regulation characteristic as shown in Figure 17 is obtained.

5.3.3 Interaction of Channel and Group Regulation

The open wire line is so equalized that all 6 transmitted carriers are of equal magnitude at the input of the group amplifier with minimum (fair-weather) line attenuation (Figure 17, case A). The controlled-gain group amplifier then has gain GA and, as will be readily seen, the output levels of all channels are the same (only

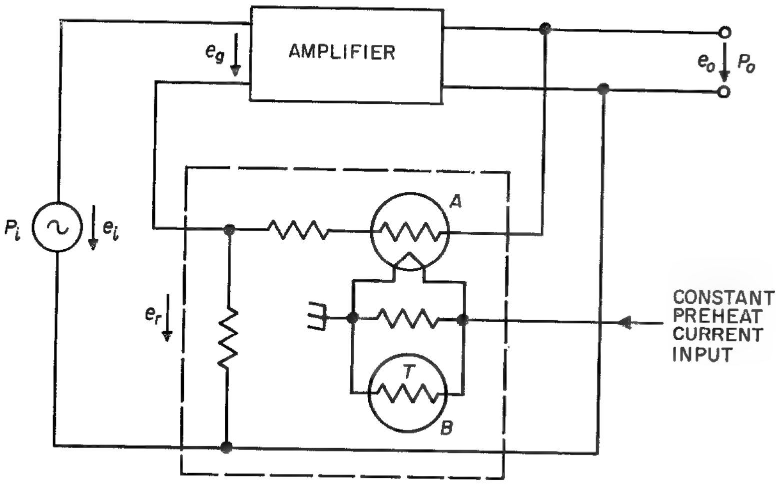
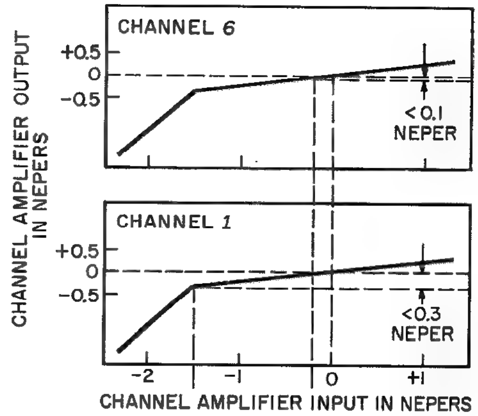


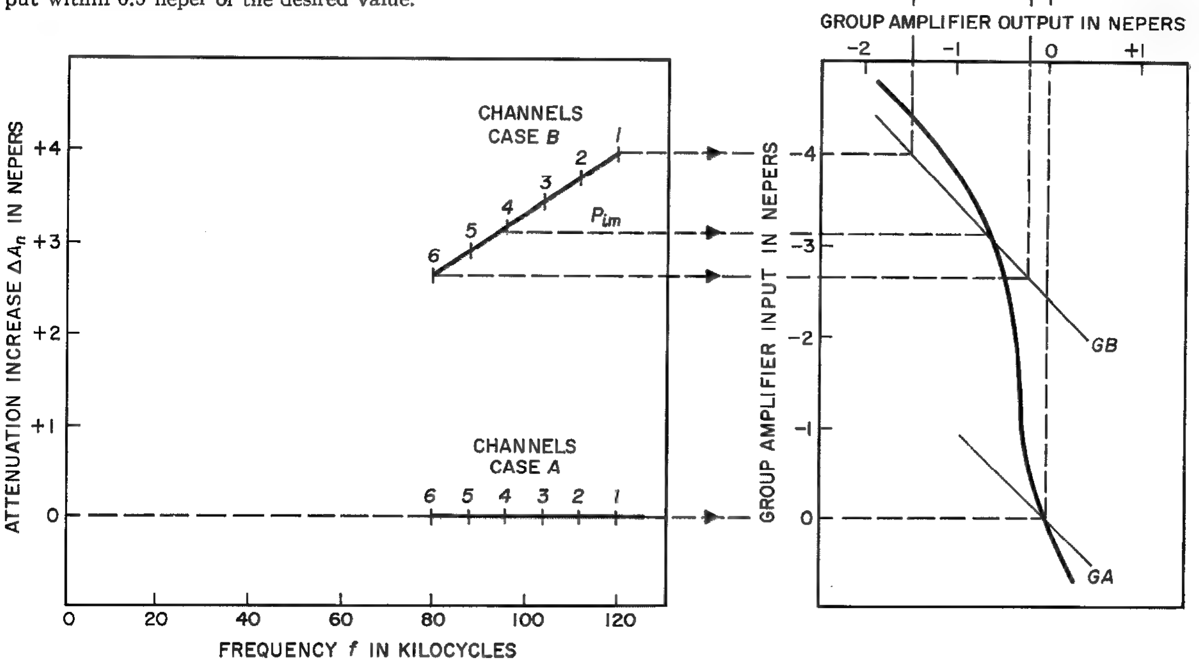
Figure 16—Basic circuit of controlled-gain group amplifier. Thermistor A determines the control characteristic and B the temperature balance. Amplification $G = (e_o/e_i) \sim (e_o/e_r)$.

Figure 17—Interaction of channel and group regulation. In the event of fair weather (case A), inputs to and outputs from group and channel amplifiers are the same for all 6 channels; the channel amplifier outputs are all P_{o-n}. In case B (foul weather), an increase in attenuation ΔA_n of 4 nepers at 120 kilocycles with a slope down to about 2.6 nepers at 80 kilocycles. The gain GB of the group amplifier in this case is determined by the median P_{im} of the levels of the 6 channels. As shown at the upper right, the equalization and regulation maintain the channel-amplifier output within 0.3 neper of the desired value.

channels *1* and *6* are shown).

Whenever weather conditions cause the open-wire-line attenuation to increase, carrier levels of different magnitudes corresponding to unequal attenuation increases appear at the group amplifier input. The limiting case B is shown in Figure 17. The gain of the group amplifier is determined by the sum of the powers of all carriers appearing at its input.





The regulation characteristic shows the dependence of output level on input level referred to one channel. Hence, the gain GB of case B is determined by the median level P_{im}. It will be seen from the diagram that the output levels of individual channel amplifiers barely change despite a large decrease of carrier level on the line. The level difference in channel *6* is under 0.1 neper; in channel *3*, about 0.3 neper.

5.3.4 Levels with Regulated Repeaters

Figure 17 refers to operation on open wire lines between terminals. When repeaters are employed having controlled gain exactly like that of the terminal gain-controlled group amplifier, the effect of group frequency translation must be taken into account.

Figure 18 shows the conditions existing when open-wire-line operation includes two repeaters. The minimum (fair-weather) line attenuation is compensated by the repeater and receiving-terminal preamplifiers. The levels shown at various points of transmission direction B-to-A refer to an attenuation in each line section of 4 nepers at 120 kilocycles. It will be seen that the original carrier level is restored after passing through two repeaters. The conditions for the A-to-B direction are analogous. The conditions prevailing in terminal A are described by the levels shown in Figure 17, case B.

Irrespective of attenuation conditions in individual sections of the open wire line, variations are always compensated as long as the levels in the last line section do not exceed the limiting case B of Figure 17.

In open-wire-line operation, the system behavior is of particular interest when rapid attenuation changes occur. Figure 19 shows some typical cases. An interesting feature is the fact

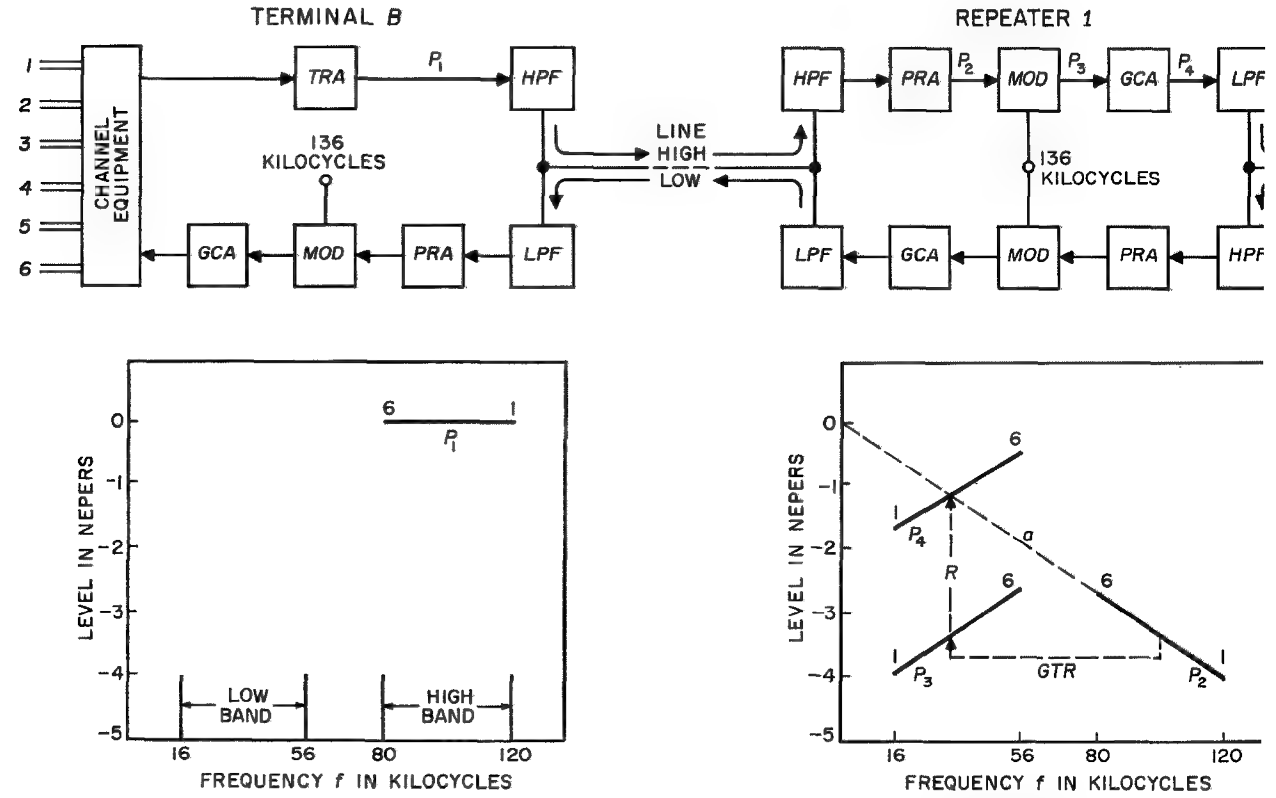


Figure 18—Level conditions with gain-controlled repeaters and group translation over open wire lines. The frequency characteristics of the line under fair-weather conditions are compensated by pads that are adjusted when the equipment is first installed but the poor-weather line-attenuation variations must be compensated by gain-controlled amplifiers in the repeaters and the receiving-terminal equipment.

In the graphs, a line attenuation versus frequency increase a having a slope of 4 nepers at 120 kilocycles under

that the process of regulation is relatively slow when attenuation increases but regulation is fast when attenuation decreases. This behavior is caused by the thermistors in the negative-feedback paths of the amplifiers.

6. *Transmission Properties*

The *Z6NC* system was developed assuming that with normal operational conditions and a line length of 100 kilometers (three sections with two repeaters), a noise power of 0.002 microwatt at reference level in a channel will be exceeded with a temporal probability of only 1 percent. The noise power was apportioned as follows.

0.001 microwatt for noise through crosstalk couplings.

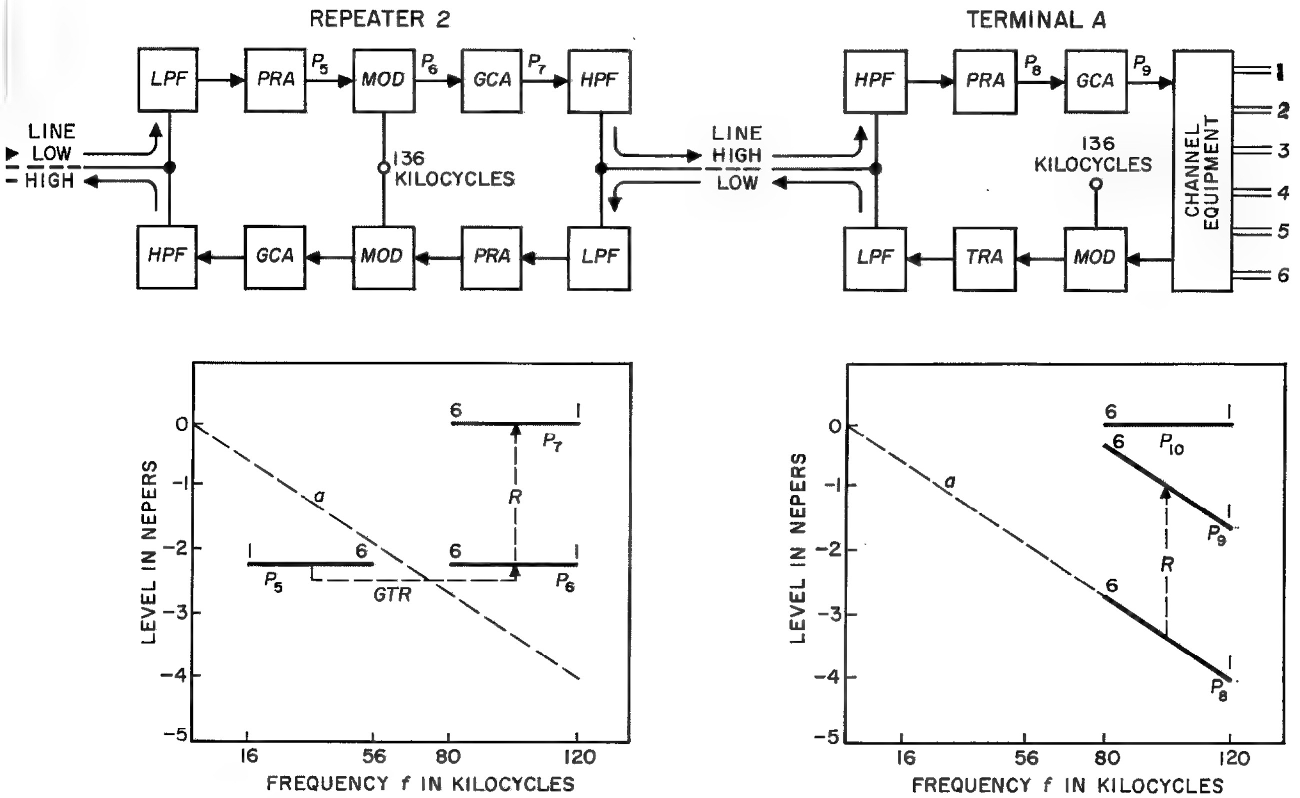
0.001 microwatt for background noise and nonlinear distortions, including here, 500 picowatts for nonlinear distortion in the transmitting and receiving circuits of a terminal.

The most-essential data of the compandor system listed below are in accordance with recommendations of the Comité Consultatif International Téléphonique.

6.1 Transmission Over Cables

6.1.1 Voice-Frequency Specifications

Transmitted band	0.3 to 3.4 kilocycles
Net loss tolerance, effective median value	⩽20 percent of Comité Consultatif International Télephonique specification



poor-weather conditions is assumed. The compensation between P_9 and channel outputs P_{10} is accomplished in the channel receiving amplifier.

TRA = transmitting amplifier, *GCA* = gain-controlled amplifier, *MOD* = modulator, *PRA* = preamplifier. *HPF* = high-pass filter, *LPF* = low-pass filter, and *GTR* = group-frequency translation and inversion between high and low bands. The compensating range of the gain-controlled amplifiers is indicated by *R*.

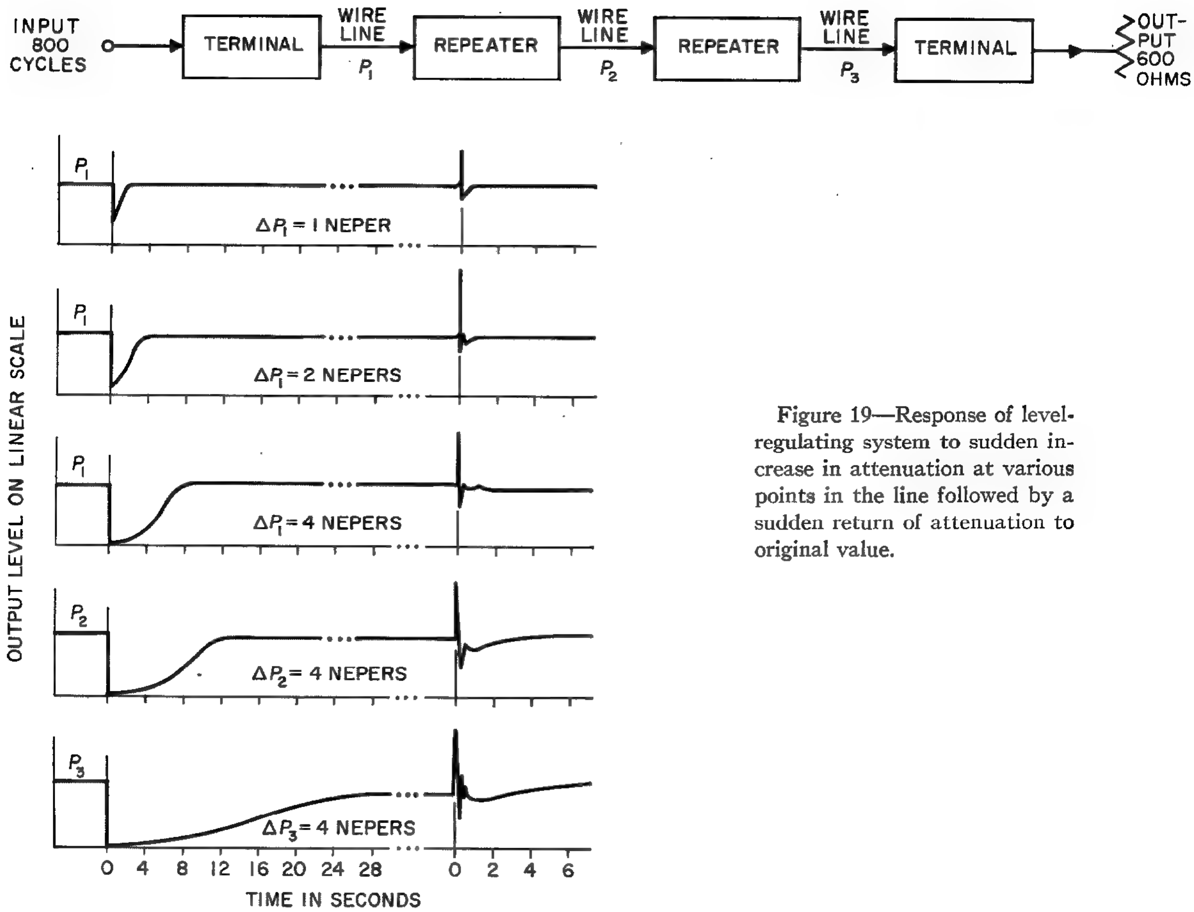


Figure 19—Response of level-regulating system to sudden increase in attenuation at various points in the line followed by a sudden return of attenuation to original value.

Total harmonic distortion at 800 cycles	
With expected level	<3 percent
With level increased by 0.7 neper	<6 percent
With level increased by 1.4 neper	<15 percent
Signal-to-intelligible-crosstalk ratio	>8.5 nepers
Signal-to-unintelligible-crosstalk ratio (buzzer tone of zero relative level imposed on any two channels)	>8 nepers
Signal-to-background-noise ratio	>8 nepers

Noise voltage in any other channel when one channel is overloaded by a 1.4-neper-level signal keyed at 50-milliseconds on, 50-milliseconds off	<1.2 millivolts*
Noise voltage in any other channel when one channel receives periodic voltage peaks of 60 volts at relative zero level (discharge of 2-microfarad capacitor through 600 ohms)	<1.2 millivolts*

* Measured at a point having a +1.0-neper relative level.

6.1.2 Signal Transmission

Signal delay time	<12 milliseconds
Signal distortion with line attenuation variations of ± 1 neper	$<\pm 1$ millisecond
Speech immunity	
Maximum overload level not generating spurious signal	$\leqslant 1.9$ nepers at relative zero level
Noise in adjacent channels with signal transmission of 50/50 or 20/20 milliseconds on/off ratio	<0.5 millivolts*
Noise voltage of continuous signal tone in the related channel	<0.6 millivolts*

6.1.3 Carrier-Frequency System

Carrier transmitter level	$+0.5$ neper (adjustable in steps of 0.1 neper to ± 0.3 neper)
Carrier transmitter level in *B* terminal can be switched to	$+1.5$ nepers
Depth of modulation on toll line (for rated level)	35 to 45 percent
Attenuation	
Between terminals	0 to 9.3 nepers (at 124 kilocycles)
With repeaters	0 to 8.3 nepers (at 124 kilocycles)
Transmission range with deloaded toll cable (1.4-millimeter copper core or 1.8-millimeter aluminum core)	
Between terminals	37 kilometers
With repeaters	33 kilometers

6.2 Transmission Over Open Wire Line

The transmission properties quoted for cable circuits are also applicable to the case of open wire lines except that the values for crosstalk are somewhat less favorable; however, this is unimportant when compared with the greater effects of noise on open wire lines.

The limits for automatic level correction due to attenuation variations are given by the conditions existing in the last section of the open wire line in the direction of transmission:

Maximum increase of attenuation	4 nepers
Maximum slope in the transmission band, for carrier-frequency levels at 120/80 and 56/16 kilocycles	1.35 nepers

The above transmission properties are maintained at ambient room temperatures of 10 to 35 degrees centigrade and 80 percent humidity. However, the equipment still operates satisfactorily in the range of 0 to 40 degrees centigrade.

Award For Planar-Grid Disc-Seal Triode

The British Royal Commission on Awards to Inventors has made a joint *ex gratia* award of the sum of £2500 for the invention of the planar-grid disc-seal triode to E. H. Ullrich of Standard Telephones and Cables Limited and his coinventors J. Foster, C. N. Smyth, and S. G. Tomlin, also of that company at the time of the invention.

The first disclosure, as far as is known, of a tube of this type took place in April 1941 when Standard Telephones and Cables Limited demonstrated to the British Admiralty a 600-megacycle-per-second radar echo amplifier incorporating a *CV16* tube, fully engineered for production to the inventors' design, that reduced the noise factor of the best receviers then known by 9 decibels and produced an average increase in service radar range of about 35 percent at that frequency. The tube was designed for grounded-grid operation with the electrodes so disposed as to form parts of coaxial-line circuits. Experimental oscillator tubes made at that time on the same principles but with suitable modifications operated near 2700 megacycles; that is, at a frequency about four times as high as had previously been achieved with triodes.

The planar-grid disc-seal triode has been further refined by numerous workers and today is used the world over for both amplification and oscillation in the ultra- and super-high-frequency bands.

Fundamental Principles of Transistors

Dr. J. Evans of Standard Telecommunications Laboratories, Enfield, has recently published a book entitled "Fundamental Principles of Transistors." It is divided into 11 chapters, 3 appendixes, and a bibliography covering the following subjects.

Chapter 1—Introduction

Chapter 2—Basic Theory of Semiconductors

Chapter 3—Measurement of Semiconductor Parameters

Chapter 4—The P-N Junction: Theory

Chapter 5—The P-N Junction: Method of Preparation

Chapter 6—Junction Transistors

Chapter 7—Point-Contact Transistors

Chapter 8—Measurement of Transistor Parameters

Chapter 9—Manufacture of Transistors

Chapter 10—Special Types of Transistors

Chapter 11—Silicon and Other Transistor Materials

Appendix 1—Teaching Transistor Physics

Appendix 2—Parameters of Some Commercial Transistors

Appendix 3—Identification of Mixed Impurities

The book is $5\frac{3}{4}$ by $8\frac{3}{4}$ inches (15 by 22 centimeters) and contains 255 pages and 140 figures. It is available from Heywood and Company, Limited, of 9 Kingsway, London, W.C.2, at 45 shillings. The book can also be obtained from D. Van Nostrand Company, 120 Alexander Street, Princeton, New Jersey, at $6.75 per copy.

Tricon, an Electric-Diagram Interlock System for Railroad Switching*

By WILHELM SCHMITZ

C. Lorenz A.G. (now Standard Elektrik Lorenz A.G.); Stuttgart, Germany

DIAGRAMMATIC interlock control systems for the operation of railroad track switches have been employed in Germany for the past decade and are designated *Dr* by the Deutsche Bundesbahn (German railroads). This designation refers to the push-button keys arranged within the diagram. The outstanding characteristic of such systems is the increasing use of electric relays. While such devices were used in the former lever-interlock systems, they depended chiefly on mechanical interlocking and couplings between lever parts, like snap switches and armature stops of switch-point levers. The conversion from mechanical to electric interconnections has introduced a remarkable degree of freedom in the arrangement of operating keys, which can now be mounted directly in a track diagram. The utilization of small uniform push-button keys permits a substantial reduction in the size of the control desks. On the other hand, the number of relays and switching circuits is increased considerably, which is to be expected.

The design of such an electrical system required not only means of economically manufacturing the extensive circuit components, but the practical requirements of planning, inspecting, and maintaining such interlock systems had to be considered. The obvious necessity of locating trouble and repairing it quickly and of being able to modify and expand installations without interrupting service suggested the subdivision of the interlock system into uniform assemblies of commonly used components with corresponding relay sets. This plan was very effective for the relatively simple tasks like the setting of switchpoints or signals. The combining of several of these actions to form a routing, however, was hardly possible with the standardized components because every railroad switch-tower has its own peculiar structural and its operational conditions. A mixed construction has resulted in which the greater part consists of standard relay sets and the smaller part is planned and constructed for the individual interlock tower. This so-called "individual wiring" still occupies 30 to 40 percent of the total circuit diagram. The new relay-operated interlock system makes use of a circuit in which the switching action for routing also is standardized. This system has the designation *Sp Dr* in the terminology of the German railroads, while the trade name is Tricon.

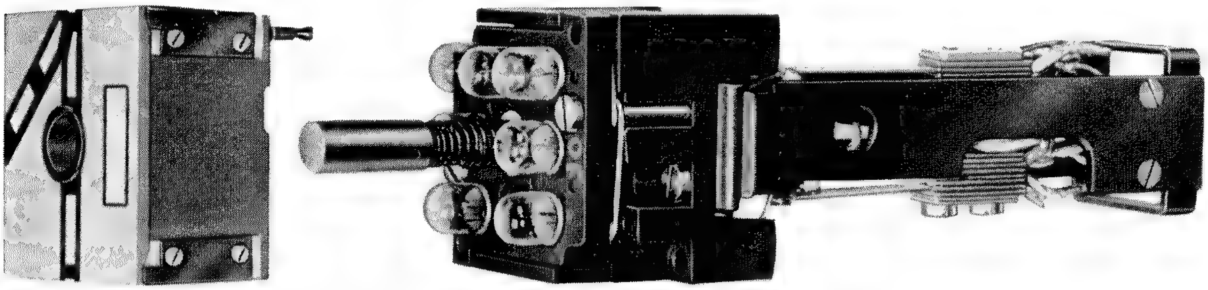


Figure 1—Track-diagram element with its cover removed. The operating push button is in the center and its contacts are in the shank of the structure. Identifying colors and graphical symbols are provided by the cover and are not affected by changing a burned-out lamp.

1. Track-Diagram Desk

The push-button keys for the operation of the interlock system are mounted in an illuminated

* Presented at a meeting of the German Railroad Engineers Society in Stuttgart on May 16, 1955. Reprinted from *Der Eisenbahningenieur*, volume 6, number 9; September, 1955.

model track diagram. To facilitate factory production and subsequent modification of any track diagram, the diagram is composed of elements not unlike the tiles in a mosaic. The elements are squares with 35-millimeter (1.38-

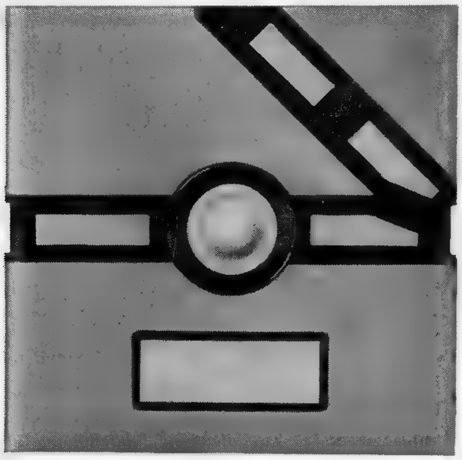


Figure 2—Top view of the cover for a single switch element. The track for which the switch is set is indicated by illumination of the corresponding graphical symbol. The identification of the switch appears in the rectangular nameplate.

inch) sides and consist of a small chassis mounting 9 electric lamps or 8 lamps with a push button mounted in the center. This chassis is screwed to longitudinal bars.

The top or cover of each mosaic element is a sheet of Plexiglas, the bottom surface of which is painted to show the desired graphical symbol of a switchpoint, signal, et cetera, illuminated as desired by electric lamps mounted in the chassis. Erroneous interchanging of colors is prevented by the use of clear glass bulbs, the colors being applied to Plexiglas caps attached to the cover. When the cover is replaced, every bulb is covered by a cap of the required color.

Figure 1 shows one of these mosaic elements with its cover removed. In Figure 2, the symbol of a switchpoint may be seen in greater detail. In the center is the push button and below it is a number plate that can be illuminated. Corresponding to each track section of the switch element are two small illuminated panels. The illumination of these two panels indicates the position of the switchpoint. While the points are being switched, a flashing light is visible on the panel corresponding to the new switch position. Occupancy of a track section by a train will be detected by insulated sections of track and will cause a red light to show in the corresponding track panels. The number plate shows a red flashing light when the switch is trailed. These mosaic elements are available with all types of symbols, such as derailers, signals, lines, and line blocks. They differ only in the symbols

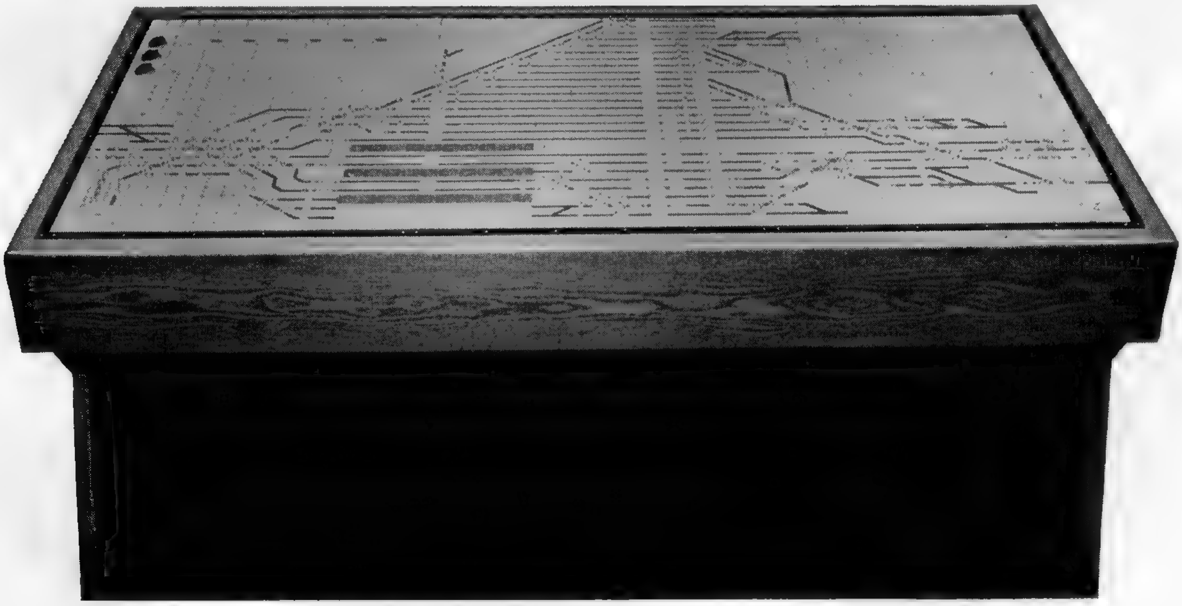


Figure 3—Typical track-diagram control desk consisting of 1512 mosaic elements.

painted on the Plexiglas covers and in the complement of lamps and colored caps.

These mosaic elements can be used to assemble track diagrams of the most diverse railroad switchtowers. Figure 3 shows a track-diagram control desk for a tower with 109 switchpoints. The number of elements making up the total picture is $28 \times 54 = 1512$, of which 385 have electrical connections. The diagram is 1890 millimeters (74 inches) long and 980 millimeters (38 inches) wide.

2. *Routing*

Figure 4 will demonstrate the principles of determining a route between two ends of a trackage system. It represents 4 tracks and 6 single switches. Based on the *N–X* principle, in the electrical routing arrangement, the pressing of any push button will make conductive every path from it through which a train can pass and will open all electrical circuits to tracks over which a train cannot pass.

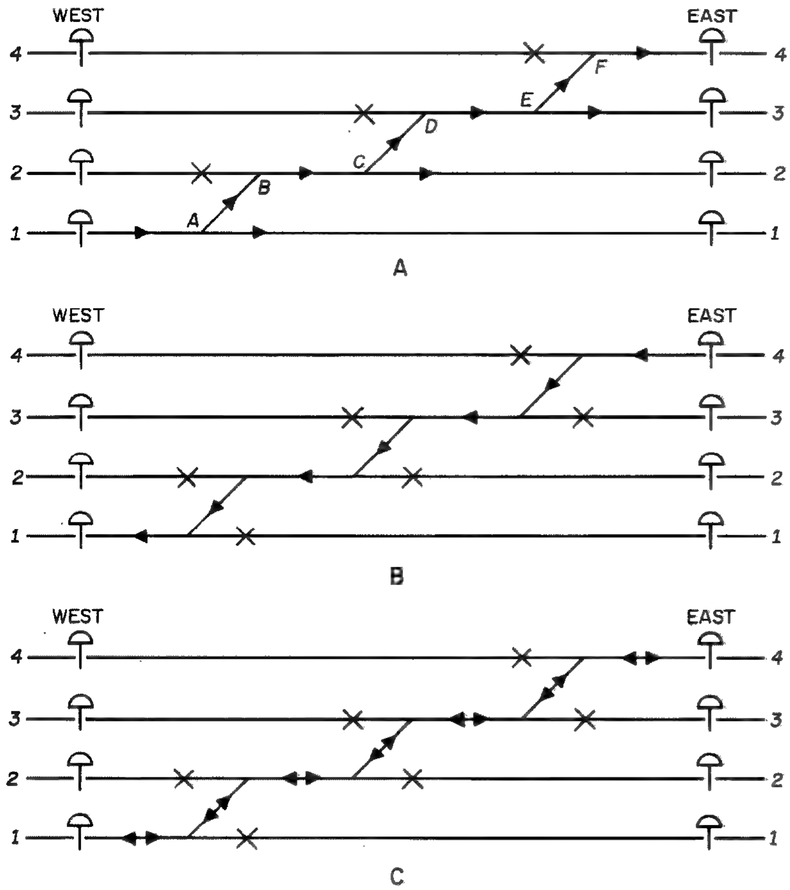


Figure 4—Arrangement of tracks and switches to illustrate route selection by pressing a push button at each end of the system. At *A*, the routes permitted are indicated by arrows and those prohibited are marked by crosses when pushbutton *1 WEST* is operated. *B* is for depressing *4 EAST* and *C* is the resultant of both being operated.

The most-common operating system of a push-button diagram desk is the *N–X* system. In it, the operating is done by pressing two buttons, one on the entrance side and the other on the exit side of the route that is to be set. In Figure 4, the route from track *4 East* to track *1 West* will be set by the operation of push buttons *4 East* and *1 West*. As shown in Figure 4A, the path finding begins at the exit point. Originating on a contact of push-button *1 West*, the pathfinding comes then to switch *A* where it finds two possible paths,one to switch *B* and the other to track *1 East*. Since the reverse (turn-out) path of switch *B* is taken, the normal (straight) way that could not be traversed by a train is cut away. In switch *C* it goes two ways, to track *2 East* and switch *D*, and so on. The pathfinding ends with access to all four *East* tracks. Since at the exit, push-button *4* is operated, the pathfinding will be reflected electrically from this point as shown in Figure 4B. At switch *E*, the normal way to track *3 East*, will be cut away, similarly in switch *C* to track *2 East*, and so on. The reflecting pathfinding ends at the originating point *1 West*. This produces an electrical channel between the entrance and the exit point, as shown in Figure 4C. Now it is not difficult to set the switches and signals as needed over this electrical channel.

3. *Shunting Route*

A shunting route will serve as an example of the operation of the system. As is well-known,

the difference between shunting and through routes is that for shunting, only the track switches over which the train will pass need be set and locked before the clear signal can be set; while for the through route, it is necessary to set switches and safety switches and to detect the entire line before the signal shows *clear*. To ensure greatest flexibility of the interlock system and to be able to release the separate route sections quickly, each switch is released as soon as its detector track is unoccupied. That means that the system retains a temporary switching route for the shortest necessary time, making the trackage available for other uses without delay.

Figure 5 shows a simple track diagram with control push buttons and shunting signals. Below the track diagram, the operating desk is seen. Below the latter are the corresponding sets of relays; that is, for 4 switches and 4 shunting signals. The sequence of operations of the relay sets are diagramed in the 8 columns underneath the corresponding switches and signals. To set a shunting movement from the *stop* condition of signal *2B* to the shunting signal *1C*, the signalman depresses the push buttons of these two signals in the desk, thus energizing the associated relays in the relay sets of signals *2B* and *1B*. It should be noted that the signal relay in relay set *1B* is energized although push button *1C* is depressed. It is necessary that an impulse be released in the relay sets associated with the points of the route switches; the two ends of this route can be controlled only by these two relays.

Now the process of routing begins. Each switch relay set has three *KS* relays. The *KS*± relay is set if the current comes from the points of the switch, *KS*+ if from the normal of the switch, and *KS*− if from the reverse side of the switch. For the example under consideration, the *KS*± relay is energized in the relay set of switch *4*. As the pulse comes from the switchpoint, a train might travel along the straight track or over the shunt or turnout route. From the point of the switch, the pulse is transmitted along the main line to switch *1*, where relay *KS*+ is energized to signify that only the straight route may be traversed as a train cannot be shunted from the reverse side of the switch. The pulse then goes to the relay set of shunting signal *1A* where a route control relay *KC* is energized. Also from the point of switch *4*, a connection to the siding of switch *3* is made where *KS* − is energized, allowing only the reverse route to be taken. *KS*+ of switch *2* will be energized, and finally route control relay *KC* in relay set *2B* is operated. Since the push-button relay *KSR1* is already energized here, another relay *KDI* is operated, determining the directional setting at this end; later this shunting signal is set to *clear*.

Now the process called echo is initiated. Switches *2* and *3* are taken facing the points and relays *KS*± are each put in the energized condition. At switch *4*, only the siding may be traversed, *KS*− blocking the main line to signal *1A* previously permitted. The echo reaches its end by energizing relay *KC* in the relay set of signal *1B*. The latter causes the directional-setting relay *KDO* to indicate that the route ends here, that is, that the shunting signal *1B*, must not be set to *clear*. Moreover, the indicator lamp is lighted in the number plate of shunting signal *1C*, informing the signalman that the desired route is electrically formed.

Now the signalman releases the push buttons so that the push-button relays are released. This causes a pulse to be transmitted from switch to switch along the channel formed by the *KS* relays, reversing or controlling the switch position relays *KP* so that they assume the position determined by the *KS* relays. If necessary, the switchpoints are controlled by the *KP* relays. When in the correct positions, all switches are locked by route locking relays *KL*. The *KL* relay next to signal relay set *1B* connects into the circuit route release relay *KR*, which is used later for the release of the route. Checking the stop signal of the shunting signal *1B*, the switch positions, and the switch locking, a current impulse is now transmitted through all switches to the shunting signal-setting relay *KIS* in the signal relay set *2B*, which sets the shunting signal to clear.

The ensuing events, beginning with the shunting movement, are represented by Figure 6. As soon as the train in the shunting movement enters the detector track section of switch *2*, the associated track relay *KT* drops. When the shunting movement advances to include switch *3* also, this track relay is also reversed. Switch *2* is then freed as the train has passed beyond it. The track relay of switch *2* is set to the *clear* position. This causes relay *KR* in the signal relay

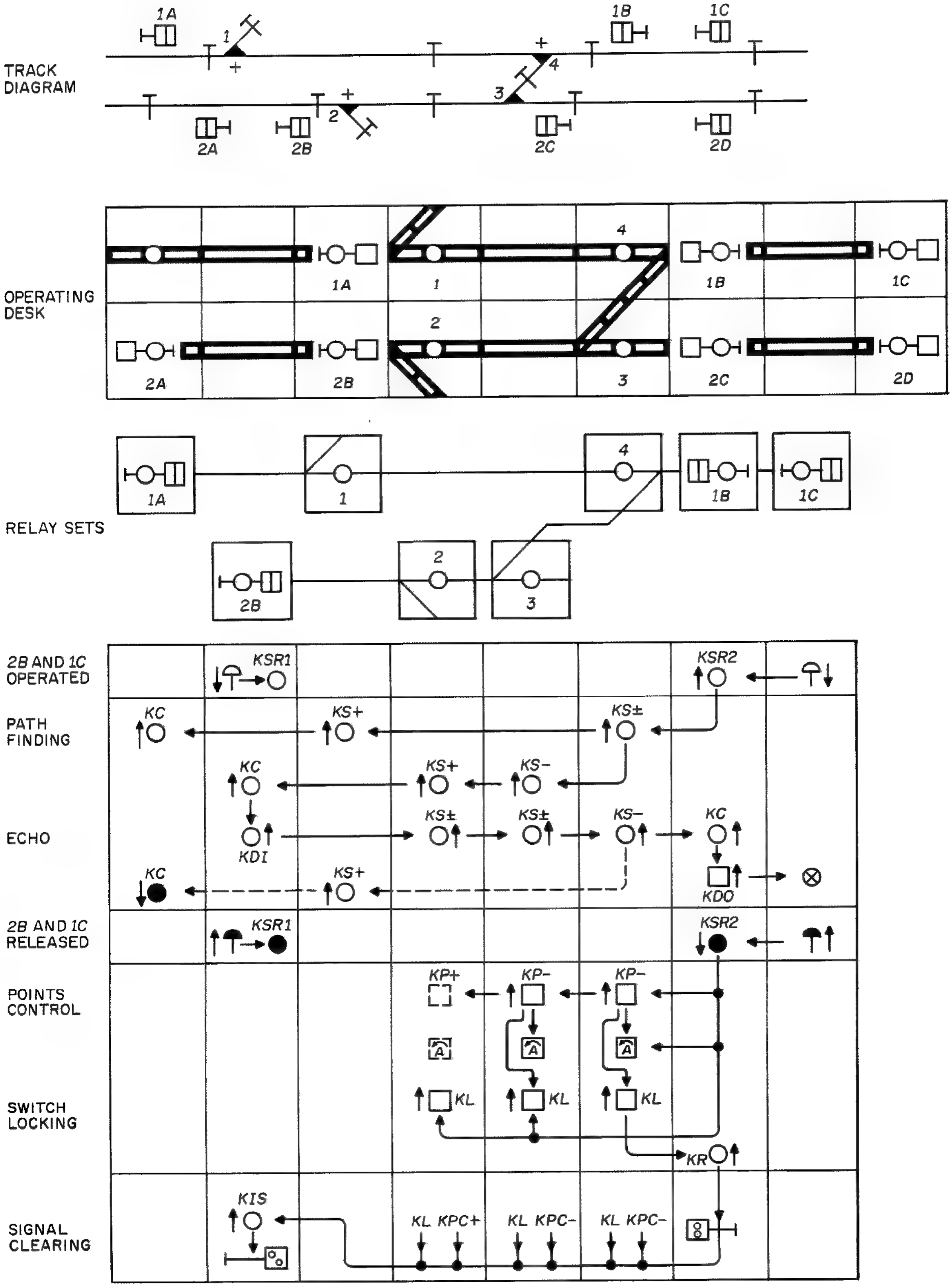


Figure 5—Track diagram, operating desk, relay sets, and sequence chart of operation to set up a shunting track connection from *2B* to *1C*. The symbols are identified in the appendix, section 7.

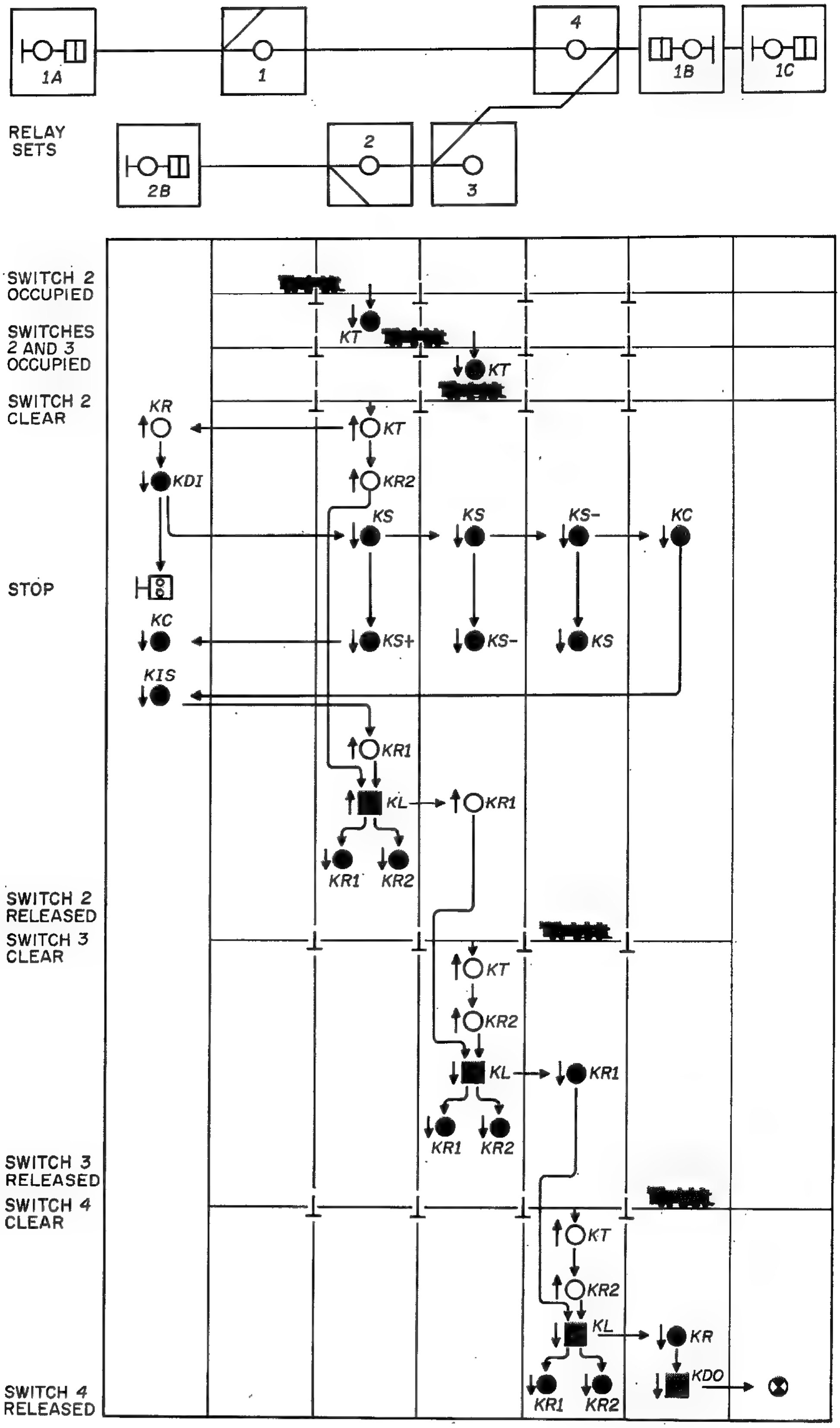


Figure 6—Release of switching elements after train has traversed the shunting route of Figure 5.

set $2B$ to be energized together with relay *KR2* in the switch relay set. Relay *KDI* in the signal relay set drops, the signal thus again assuming the *stop* position. Moreover, *KDI* releases sequentially the chain of *KS* relays for the switches,

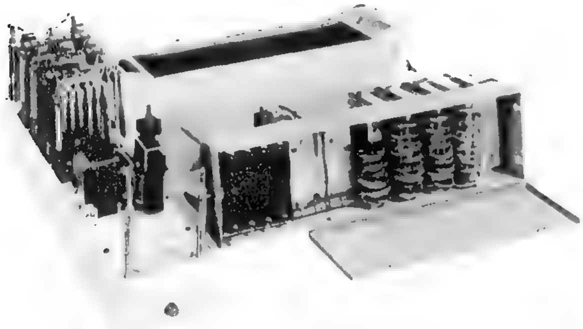


Figure 7—Typical relay arrangement.

which release the *KC* relays in both signal relay sets, and relay *KIS* drops back to its original position. This causes relay *KR1* of switch *2* to be operated. Both relays *KR1* and *KR2* reverse the position of holding relay *KL*, which disconnects relays *KR1* and *KR2* and energizes relay *KR1* of switch *3*. When the train in the shunting movement runs over the detector track circuit of switch *4*, its track relay is set for the occupied condition. As soon as the detector track section of switch *3* is freed, this switch is released in the same way as described before. The same process repeats itself when the detector track section of switch *4* is freed; through reversal of the switch locking relay, relays *KR* in the signal relay set of the route end as well as relay *KDO* is released to normal

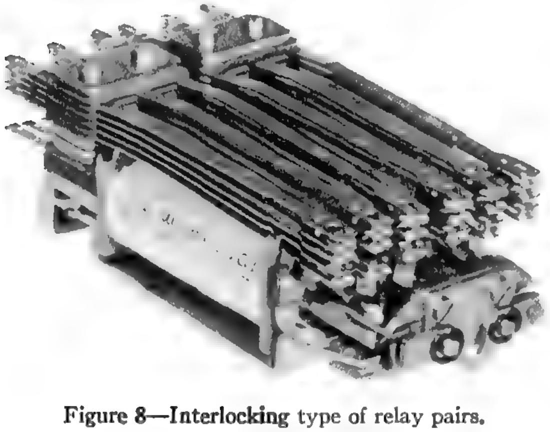


Figure 8—Interlocking type of relay pairs.

position so that the key indicator lamp of the route-end or exit push button is switched off.

It will be clear now that the switching operation proceeds from one relay set to the next set on the track diagram without any connections needed among relay sets that are not adjacent to each other on the track diagram. The same procedure is also used for through movements, in which cases it is supplemented by flank protection and clear-route detection.

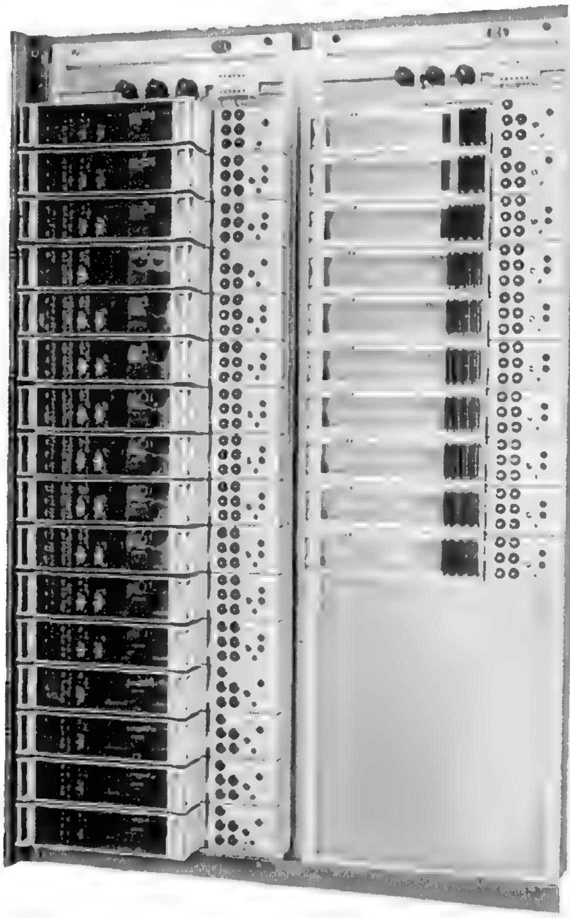


Figure 9—Relay rack with fuses and signal lamps at the right.

4. Constructional Details

The relays are mounted on panels, the connections to which are made through terminal strips. Figure 7 shows a set of relays that are mounted with other switching components on a chassis. At the rear are the wiring and the terminal strips. The wiring is covered by the plate visible in front of the chassis at the right.

Behind this chassis is the protecting hood with a Plexiglas window. A relay set of the same type is shown to the left.

Interlocked relays are shown in Figure 8. They consist of two relays, the armatures of which are mechanically coupled by locking pawls. The release of either armature will provide mechanical support for the operation of the other armature.

A complete relay rack is shown in Figure 9, the relays having been removed from the right half of the rack. To the right of each relay set is a switchboard mounting fuses and lamps according to special requirements for that particular relay set. The lamps indicate the same operating states as are indicated at the control desk and include one lamp each for normal control, reverse control, switch occupancy, and switch locking.

Figure 10 is a schematic circuit diagram for two track switches. The track-switch symbols at the top correspond to two associated relay sets, each with three connections for the point and the two paths of a switch. Each switch relay set is connected to a corresponding mosaic element of the control desk. Every two switch relay sets are connected to a common relay set connected to two insulated rails to detect the presence of a train that would provide an electric circuit across the rails. Cables lead from these relay sets to the insulated rails and to the switch machines. The designation of this interlock system, TRICON, is derived from the TRIple CONnection of the switch relay set that makes it suitable for any track diagram. It should be noted that a route in its outmoded sense is no longer in existence and that no route relay set is needed. Each relay set for switches, signals, et cetera, comprises all components required for a route. Each switch, as long as it has its own detector track circuit, is released separately and is equipped with all circuit components that are required. For instance, each switch has its own arrangements for flank protection, regardless of whether they are needed or not.

5. *Outdoor Equipment*

The signal lights used are of the standard type employed by the German railroads.

The tracks circuits are supplied with alternating current of 220 volts. The voltage is stepped down by a transformer at the supply end and stepped up in the interlock tower at the relay end, where it is applied to the grid of a vacuum tube to control two relays alternately. If the normal–clear relay is operated and the occupancy relay is released, it indicates that the track circuit is not occupied. If, on the other hand, the normal–clear relay is released and the occupancy relay operated, the track circuit is occupied. If both relays are operated or released simultaneously, this indicates trouble.

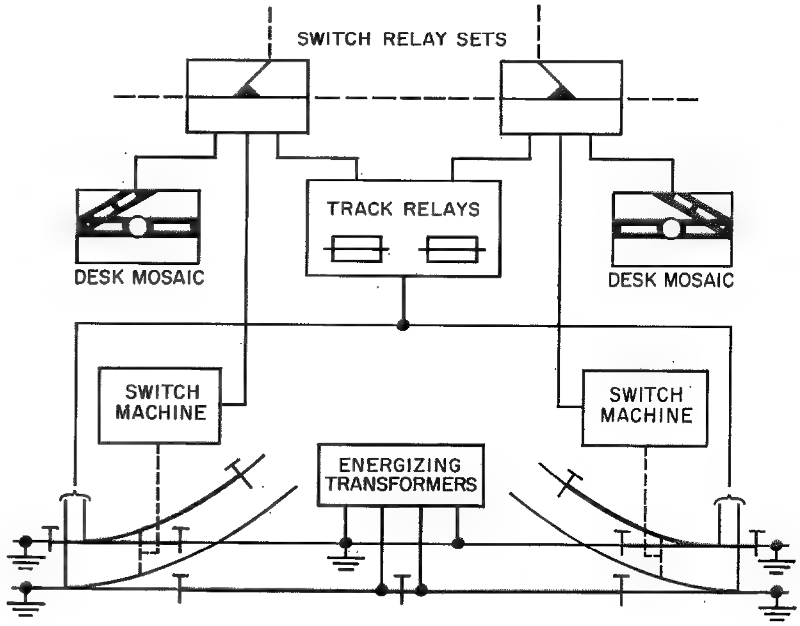


Figure 10—Schematic arrangement of switching system. Heavy lines indicate insulated sections of track, all others being grounded.

A new type of switch machine is shown in Figure 11. Through a coupling, the motor drives a speed-reducing pair of gears having a worm gear on the output shaft. The worm gear engages a toothed segment that operates an arm below the housing of the drive. This arm is coupled via rodding to the track switch. The control contacts

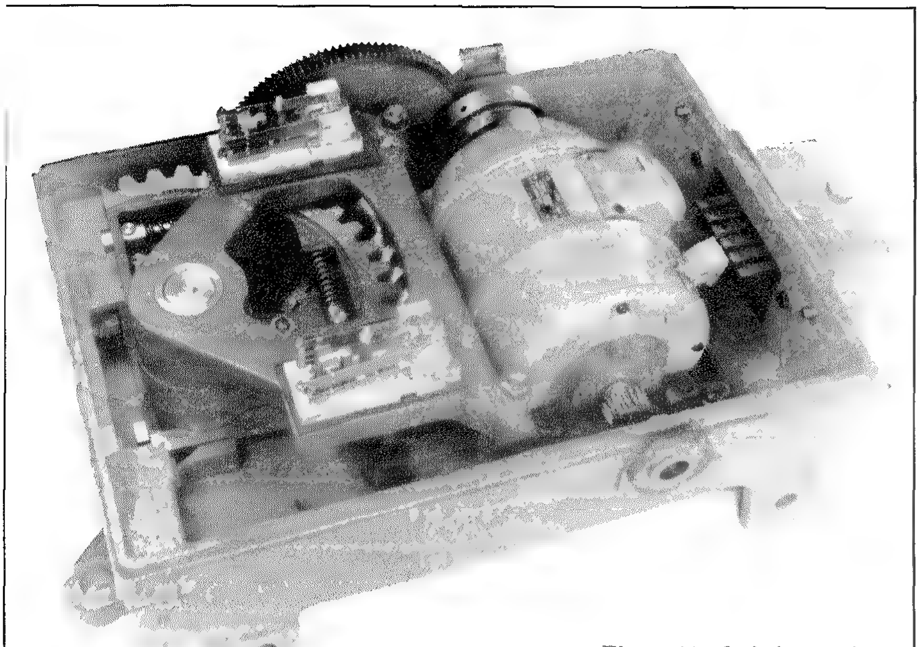


Figure 11—Switch machine.

are mounted on a large casting that supports the main bearing of the drive. The point-detector slides are mounted in the drive housing to the left. A circular opening visible in the front side of the case permits the insertion of a cranking handle acting on the motor shaft through a pinion gear. When the handle is inserted, the machine is electrically disconnected by a lever.

6. *Advantages of Tricon*

Except for the wiring diagrams for the relay sets, no other diagrams are needed at the interlock towers. The wiring is totally standardized. This enables data for orders to be processed in the workshop much sooner than for custom designs. In manufacture, there is a noteworthy advantage in the omission of individual wiring for single-interlock towers, which previously made up a considerable portion of the work. The mounting is greatly simplified. Since the connections between relay sets have to be wired rather schematically according to the track diagram, they are soldered to prenumbered cables on the construction site.

Particular advantages have been noted in the process of inspecting and placing a completed interlock system in service. A new testing method is being developed for the system.

The greatest advantage, however, will become apparent when interlock systems in operation have to be enlarged at a later date. If, for instance, a new switch with a siding is to be added in a route of switchpoints already in existence, the new siding being provided for through traffic also, then only the connection between the existing adjacent switch relay sets need be broken to insert a new factory-wired relay set together with a signal relay set for the departure signal; these two sets are then connected by indoor cables according to the track diagram, and the new switchpoint is automatically connected to all routes. The new signal is properly inserted with respect to all existing switch relay sets, including the blocking of all conflicting routes.

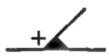
Even if the circuit investment is in some cases greater than for other systems, in which switches do not serve as protecting switches, this interlock system offers so many advantages that the slightly greater investment is easily outweighed. This system will enable railroads to complete the construction of interlock towers promptly and will accelerate the movement of rolling stock.

7. *Appendix*

The graphical symbols used in the drawings and the sequence-diagram symbols are defined in this section. It is supposed that readers of *Electrical Communication* may not be familiar with railway symbols.

7.1 Graphical Symbols

Track switch normally set for straight normal route.



Track switch normally set for reverse route.

 Insulated rail joint to detect presence of car.

 Signal post.

 Sign for shunting signal.

○ Push button.

 Signal post, push button for the shunting signal, and signal sign. Push button may be operated as the entrance or exit point for setting up a shunting route.

7.2 Sequence-Diagram Symbols

 Push buttons in open- and closed-circuit conditions, respectively.

 Push-button indicator lamps, dark and illuminated, respectively.

○ ● Relays in up and down conditions, respectively.

□ ■ Interlocked relays in released and operated conditions, respectively.

 Shunting signal lamps displaying red and white signals, respectively.

Switch machine. Arrow indicates operation to reverse position.

→ Direction of circuit action or of traffic.

↑ Direction of initiated operation. Actuation of relay or release of push button.

↓ Direction of initiated operation. Release of relay or depression of push button.

\+ Condition of relay to control track switch for for straight route.

− Condition of relay to control track switch for reverse position route.

± Relay to control track switch for either straight or turnout route. This relay operates if the route search approaches a track switch from its point side.

Loss Formulas for Homogeneous Gradings of the Second Order in Telephone Switching Employing Random Hunting

By JAKOB KRUITHOF

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PROBABILITIES for the occurrence of specific traffic patterns and consequently the equation stating the grade of service rendered by a certain grading arrangement can in principle be established by the application of the well-known theory advanced by Erlang. This theory of the so-called "statistical equilibrium" is based on the fact that for random traffic the various probabilities for the occurrence of specific traffic patterns do not depend on time. It leads to a system of equations of the type

$$(sy + c)\, p[c] = \sum y\, f(c)\, p[c-1] + \sum F(c)\, p[c+1] \quad (1)$$

where

y = average traffic offered to each split of the grading

s = number of multiple splits

c = number of calls in progress

$p[c]$ = probability of occurrence of a certain call pattern comprising c calls

$p[c-1]$ = probability of occurrence of derived patterns having one call fewer

$p[c+1]$ = probability of occurrence of derived call patterns having one more call

$f(c)$, $F(c)$ = numerical functions of c, depending on the type of grading chosen and the mode of hunting, but independent of y.

The application of this method to the computation of the expressions for the different probabilities appears to be a laborious task, especially for gradings of appreciable size as the equations rapidly reach a great number.

Since it was demonstrated by extensive and systematic tests with the rotary traffic machine[1] that gradings based on outlets connected to the same number of splits in a cyclic manner rendered improved efficiency, and since it has been shown in practice that such gradings are less sensitive to an unbalanced load, they have aroused increasing interest both on the part of the telephone administrations and the manufacturers of switching equipment.

These gradings, which we termed "homogeneous gradings," have become of great interest compared with gradings of a mixed order, as their efficiency does not depend appreciably on the mode of hunting. With several modern telephone switching systems, the hunting of a selector starts from a random point; with others, the switches are provided with a home position and hunting always commences from the same point from which the outlets are searched in a definite sequential manner.

Tests with the traffic machine have demonstrated that gradings of mixed order have poor efficiency when subjected to random hunting. Such gradings are, therefore, unsuitable for systems using nonhoming switches, while homogeneous gradings are apposite.

This paper deals exclusively with homogeneous gradings of the second order that are composed of complete cyclic arrangements of groups of outlets. It is desirable to develop exact methods that do not require the solving of an extensive system of equations.

Figure 1 is a general illustration of the homogeneous type of grading.

A symmetrically loaded, complete, homogeneous grading of the second order is fully defined by its number of splits and its number of outlets per subgroup. A subgroup includes those outlets that are common to any two specific multiple splits. The number of subgroups, therefore,

[1] J. Kruithof, "Rotary Traffic Machine," *Electrical Communication*, volume 23, pages 192–211; June, 1946.

amounts to $\binom{s}{2}$. Further,

$$N = n\binom{s}{2}$$
$$a = n(s-1)$$

where

N = total number of outlets to which the grading provides access

n = number of outlets of one subgroup

a = accessibility or availability, that is, the number of outlets to which the switches of a split provide access.

The accessibility a is the same for all splits.

The traffic offered to the various splits is assumed equal, as balanced traffic distribution is a principal aim when planning telephone exchange equipment. Unsymmetrical traffic distribution, although occurring in practice, remains an undesirable exception.

1. Examination of Probabilities

The solution of the system of (1) provides expressions for the various individual probabilities p relating to specific and definite traffic patterns. It appears that each of these can be written as a product of three factors: $p(o)$, $(sy)^i : i!$, and a fraction of which both the numerator and the denominator are polynomials in y and of equal degree.

For the sum P of the probabilities relating to the same number of simultaneous calls, the following types of expressions are found.

$$\left.\begin{aligned} P(1) &= (sy)\,P(o)\\ P(2) &= \frac{(sy)^2}{2!}P(o)\\ &\cdots\\ P(a) &= \frac{(sy)^a}{a!}P(o). \end{aligned}\right\} \quad (2A)$$

$$\left.\begin{aligned} P(a+1) &= k_{a+1}\frac{(sy)^{a+1}}{(a+1)!}P(o)\\ P(a+2) &= k_{a+2}\frac{(sy)^{a+2}}{(a+2)!}P(o)\\ &\cdots\\ P(N) &= k_N\frac{(sy)^N}{N!}P(o). \end{aligned}\right\} \quad (2B)$$

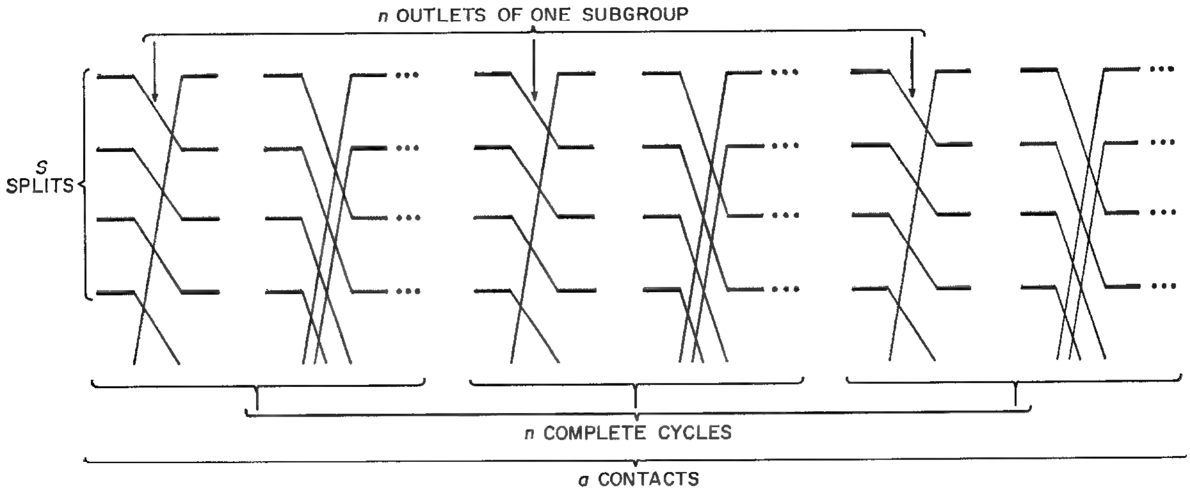


Figure 1—General illustration of homogeneous grading.

That is, after the summation of the p's, the sums of the fractional factors appear to be equal to unity for values of $i \leqslant a$, as could be expected. For the probabilities P for traffic patterns having more than a simultaneous calls, a coefficient $k < 1$ remains.[2]

[2] M. Van den Bossche and J. Kruithof, "Some Notes on the Mathematical Treatment of Homogeneous Gradings," Memorandum MSW 83, International Telephone and Telegraph Corporation. In this office memorandum, the writers raise several objections against the mathematical treatment of the grading problem developed in the paper by A. Elldin, "On the Congestion in Gradings with Random Hunting," that appeared in *Ericsson Technics*, number 1; 1955. Section 1 of the present paper repeats their point of view as regards the correct reasons and conditions why and when (1) may be separated into *two* incompatible systems of equations.

These coefficients k are similar to those mentioned for the p's and consist of a fraction of which both the numerator and the denominator are polynomials in y and of equal degree. The expressions for these coefficients depend on the type of grading chosen, the mode of hunting, and the traffic offered to the grading.

A few arbitrarily chosen coefficients follow.

In the exact solution of the probabilities $P(5)$ and $P(6)$ for a three-split homogeneous grading having two outlets per subgroup ($s=3$, $n=2$) the coefficients k include the fraction

$$A = \frac{6561\,y^3 + 19\,197\,y^2 + 20\,628\,y + 7900}{297\,y^3 + 871\,y^2 + 938\,y + 360}$$

where y is the average traffic offered to each split.

In the exact solution for a four-split homogeneous grading having one outlet per subgroup ($s = 4$, $n = 1$), the probability $P(4)$ of finding any call pattern with four simultaneous calls, includes the coefficient

$$B = \frac{43\,776\,y^4 + 128\,784\,y^3 + 146\,668\,y^2 + 77\,421\,y + 16\,080}{46\,848\,y^4 + 138\,000\,y^3 + 157\,344\,y^2 + 83\,136\,y + 17\,280}$$

while the coefficients of $P(5)$ and $P(6)$ contain the fraction

$$C = \frac{34\,560\,y^4 + 101\,520\,y^3 + 115\,440\,y^2 + 60\,840\,y + 12\,615}{46\,848\,y^4 + 138\,000\,y^3 + 157\,344\,y^2 + 83\,136\,y + 17\,280}.$$

It will be agreed that, even for the above two simple examples of gradings, the coefficients have a somewhat complicated appearance.

Similar coefficients are always found in conjunction with probability problems of limited accessibility. Their values vary between two limits determined[3] by $y = 0$ and $y = \infty$.

The coefficients, however, that relate to the homogeneous gradings that are dealt with in this paper show a remarkable property. It appears that the limiting values for the coefficients lie very close together. Table 1 gives the limiting values for the above three fractions.

The difference between the corresponding values of the above two columns appears not to exceed approximately 1 percent. This convergence increases with the values of s and n and permits considerable simplification of (1).

[3] H. A. Longley, "Efficiency of Gradings," *Post Office Electrical Engineers Journal,* volume 41; April and July, 1948.

As already stated, the probabilities $p(c)$ relating to specific traffic patterns having c simultaneous calls consists of a factor y^c and a coefficient similar to those of the P's. Consequently (1) includes two types of terms, the one type having y^c as a factor and the other y^{c+1}. For $y = 0$, the terms including y^{c+1} disappear; and, for $y = \infty$, the terms containing y^c disappear. Thus the following two incompatible systems of recurrent equations[2] are obtained.

$$c\,p[c] = \sum y\,f(c)\,p[c-1], \quad \lim y \to 0. \tag{3A}$$

$$sy\,p[c] = \sum F(c)\,p[c+1], \quad \lim y \to \infty. \tag{3B}$$

The solution of each of these two systems of equations provide two limiting values for the probabilities p relating to specific call patterns.

Either system may be used for obtaining a very close approximation of the wanted probabilities.

In the following, use will be made of only the system of equations valid for $y \to \infty$ because it provides more-convenient results and also includes a small safety margin.

For symmetrically loaded homogeneous gradings consisting of twos exclusively, the complete system of equations for (3B) is

$$\begin{aligned} sy\,p[{}_1c_2,\, {}_2c_3 \cdots {}_1c_3,\, {}_2c_4 \cdots] &= ({}_1c_2 + 1)\,p[{}_1c_2 + 1,\, {}_2c_3 \cdots {}_1c_3,\, {}_2c_4 \cdots] \\ &+ ({}_2c_3 + 1)\,p[{}_1c_2,\, {}_2c_3 + 1 \cdots {}_1c_3,\, {}_2c_4 \cdots] \\ &+ \cdots \end{aligned} \tag{4}$$

The suffixes relate to the splits. Therefore, ${}_1c_2$ indicates the number of busy outlets of the 1–2

TABLE 1

LIMITING VALUES OF COEFFICIENTS

Coefficient	$y=0$	$y=\infty$
A	21.9444	22.0909
B	0.9304	0.9344
C	0.7300	0.7377

subgroup, that is, the number of the outlets common to splits 1 and 2.

The sum of the ${}_ic_j$ for all subgroups is equal to c.

The number of c's appearing in the above equation is equal to the number of subgroups, that is, $\binom{s}{2}$.

2. *General Loss Equation*

The lost traffic equals the traffic offered to the grading reduced by the traffic carried by the connected outlets.

$$sy - [P(1) + 2\,P(2) + 3\,P(3) + \cdots (N-1)\,P\,(N-1) + NP(N)].$$

Introducing the values of (2), there is obtained for the grade of service[2]

$$W = \frac{(1-k_{a+1})\dfrac{(sy)^a}{a!} + (k_{a+1} - k_{a+2})\dfrac{(sy)^{a+1}}{(a+1)!} + \cdots + k_N \dfrac{(sy)^N}{N!}}{1 + sy + \dfrac{(sy)^2}{2!} + \cdots + \dfrac{(sy)^a}{a!} + k_{a+1}\dfrac{(sy)^{a+1}}{(a+1)!} + \cdots + k_N \dfrac{(sy)^N}{N!}}. \qquad (5)$$

As a point of later importance, it should be noted that the relation given in (6) exists between the coefficients of the terms appearing in the numerator.

$$(1 - k_{a+1}) + (k_{a+1} - k_{a+2}) + \cdots k_N = 1. \qquad (6)$$

The polynomial of the numerator consists of a total of $(N + 1 - a) = n\binom{s-1}{2} + 1$ terms; the denominator has $(N + 1) = n\binom{s}{2} + 1$ terms, the first $(a + 1) = n(s - 1) + 1$ of which have a coefficient 1.

The number of coefficients amounts to $(N - a) = n\binom{s-1}{2}$.

3. *Loss Equations for a Few Specific Homogeneous Gradings* ($y \to \infty$)

Starting from $s = 2$, that is, an ideal group, we have calculated in accordance with (4) the equations for the grade of service for a number of cases that employ complete cyclic arrangements. For the $s = 2$ cases, the B equation of Erlang applies.

$n=1$

$$\left.\begin{matrix} s=2 \\ a=1 \end{matrix}\right\} \frac{(2y)}{1+(2y)} \quad \text{(Erlang)}.$$

$$\left.\begin{matrix} s=3 \\ a=2 \end{matrix}\right\} \frac{\dfrac{(3y)^2}{2!}[1+(2y)]}{3\left[1+(3y)+\dfrac{(3y)^2}{2!}+\dfrac{2}{3}\dfrac{(3y)^3}{3!}\right]}.$$

$$\left.\begin{matrix} s=4 \\ a=3 \end{matrix}\right\} \frac{\dfrac{(4y)^3}{3!}\left[\sum_0^2 \dfrac{(3y)^i}{i!}+\dfrac{2}{3}\dfrac{(3y)^3}{3!}\right]}{\dfrac{61}{4}\left[\sum_0^3 \dfrac{(4y)^i}{i!}+\dfrac{57}{61}\dfrac{(4y)^4}{4!}+\dfrac{45}{61}\dfrac{(4y)^5}{5!}+\dfrac{45}{122}\dfrac{(4y)^6}{6!}\right]}.$$

$$\left.\begin{matrix} s=5 \\ a=4 \end{matrix}\right\} \frac{\dfrac{(5y)^4}{4!}\left[\sum_0^3 \dfrac{(4y)^i}{i!}+\dfrac{57}{61}\dfrac{(4y)^4}{4!}+\dfrac{45}{61}\dfrac{(4y)^5}{5!}+\dfrac{45}{122}\dfrac{(4y)^6}{6!}\right]}{\dfrac{33\,573}{305}\left[\sum_0^4 \dfrac{(5y)^i}{i!}+\dfrac{33\,268}{33\,573}\dfrac{(5y)^5}{5!}+\dfrac{32\,048}{33\,573}\dfrac{(5y)^6}{6!}+\dfrac{29\,120}{33\,573}\dfrac{(5y)^7}{7!}+\dfrac{39\,424}{55\,955}\dfrac{(5y)^8}{8!}+\dfrac{129\,024}{279\,775}\dfrac{(5y)^9}{9!}+\dfrac{258\,048}{1\,398\,875}\dfrac{(5y)^{10}}{10!}\right]}.$$

$n=2$

$$\left.\begin{matrix} s=2 \\ a=2 \end{matrix}\right\} \frac{\frac{(2y)^2}{2!}}{1+(2y)+\frac{(2y)^2}{2!}} \quad \text{(Erlang)}.$$

$$\left.\begin{matrix} s=3 \\ a=4 \end{matrix}\right\} \frac{\frac{(3y)^4}{4!}\sum_0^2 \frac{(2y)^i}{i!}}{11\left[\sum_0^4 \frac{(3y)^i}{i!}+\frac{10}{11}\frac{(3y)^5}{5!}+\frac{20}{33}\frac{(3y)^6}{6!}\right]}.$$

$$\left.\begin{matrix} s=4 \\ a=6 \end{matrix}\right\} \frac{\frac{(4y)^6}{6!}\left[\sum_0^4 \frac{(3y)^i}{i!}+\frac{10}{11}\frac{(3y)^5}{5!}+\frac{20}{33}\frac{(3y)^6}{6!}\right]}{\frac{10\,343}{32}\left[\sum_0^6 \frac{(4y)^i}{i!}+\frac{10\,311}{10\,343}\frac{(4y)^7}{7!}+\frac{10\,143}{10\,343}\frac{(4y)^8}{8!}+\frac{9\,639}{10\,343}\frac{(4y)^9}{9!}+\frac{8\,505}{10\,343}\frac{(4y)^{10}}{10!}+\frac{25\,515}{41\,372}\frac{(4y)^{11}}{11!}+\frac{25\,515}{82\,744}\frac{(4y)^{12}}{12!}\right]}.$$

$n=3$

$$\left.\begin{matrix} s=2 \\ a=3 \end{matrix}\right\} \frac{\frac{(2y)^3}{3!}}{1+(2y)+\frac{(2y)^2}{2!}+\frac{(2y)^3}{3!}} \quad \text{(Erlang)}.$$

$$\left.\begin{matrix} s=3 \\ a=6 \end{matrix}\right\} \frac{\frac{(3y)^6}{6!}\sum_0^3 \frac{(2y)^i}{i!}}{43\left[\sum_0^6 \frac{(3y)^i}{i!}+\frac{42}{43}\frac{(3y)^7}{7!}+\frac{112}{129}\frac{(3y)^8}{8!}+\frac{224}{387}\frac{(3y)^9}{9!}\right]}.$$

$n=4$

$$\left.\begin{matrix} s=2 \\ a=4 \end{matrix}\right\} \frac{\frac{(2y)^4}{4!}}{\sum_0^4 \frac{(2y)^i}{i!}} \quad \text{(Erlang)}.$$

$$\left.\begin{matrix} s=3 \\ a=8 \end{matrix}\right\} \frac{\frac{(3y)^8}{8!}\sum_0^4 \frac{(2y)^i}{i!}}{\frac{521}{3}\left[\sum_0^8 \frac{(3y)^i}{i!}+\frac{518}{521}\frac{(3y)^9}{9!}+\frac{500}{521}\frac{(3y)^{10}}{10!}+\frac{440}{521}\frac{(3y)^{11}}{11!}+\frac{880}{1563}\frac{(3y)^{12}}{12!}\right]}.$$

4. Conclusions

It appears from the calculations in section 3, that the numerators as well as the denominators consist of two factors.

A numerator always includes the factor $(sy)^a\!:a!$ and a polynomial. A denominator always includes a polynomial and a factor K. For $s = 2$, the polynomial of the numerator and the factor K are equal to 1.

It will be noted that a relation exists between the two equations corresponding to consecutive gradings. The polynomial of the denominator of a grading determined by $n = n_1$ and $s = s_1$ corresponds to the polynomial of the numerator of the grading determined by $n = n_1$ and $s = s_1 + 1$.

This phenomenon is general and provides a key to the equation for any wanted homogeneous grading. From the Erlang formula for a grading $(n, s = 2)$, the equation for the grading $(n, s = 3)$ is derived. From the latter, that for the grading $(n, s = 4)$ can be found, et cetera.

Once the polynomial of the numerator of the wanted equation is known, the calculation of the factor K follows from (6) and, from the product of this polynomial, and the factor $(sy)^a:a!$. By simple deduction, the factor K is equal to a polynomial similar to the one found for the numerator but in which the factors $(s-1)y$, $[(s-1)y]^2:2!$, $[(s-1)y]^3:3!$, et cetera are replaced by

$$\frac{s-1}{s}\binom{a+1}{1}, \quad \left(\frac{s-1}{s}\right)^2\binom{a+2}{2},$$
$$\left(\frac{s-1}{s}\right)^3\binom{a+3}{3}, \quad \text{et cetera.}$$

The coefficients k_{a+1}, k_{a+2}, et cetera, of the polynomial of the denominator can be derived from the found coefficients of the numerator, that is, from $(1-k_{a+1})$, $(k_{a+1}-k_{a+2})$, et cetera.

That a relation exists between the equations of consecutive gradings also appears from the fact that a grading determined by $n = n$, and $s = s_1 + 1$ and in which all outlets of one split happen to be occupied, behaves like a grading determined by $n = n$, and $s = s_1$. The average traffic offered to each split equals y for both cases.

As a conclusion, it can be stated that the loss equation for any wanted, complete, homogeneous grading of the second order and for the limit $y \rightarrow \infty$ can be found by the application of a recurrent method. The B formula of Erlang serves as a starting point.

5. *Application*

To calculate the equation indicating the grade of service for a homogeneous grading typified by $(n = 3, s = 4)$, the equation for the grading $(n = 3, s = 3)$ is used as a starting point (see section 3).

The polynomial of the numerator of the wanted equation can be written immediately as

$$\sum_0^6 (3y)^i:i + \frac{42}{43}\frac{(3y)^7}{7!} + \frac{112}{129}\frac{(3y)^8}{8!} + \frac{224}{387}\frac{(3y)^9}{9!}.$$

The numerator further includes the factor $(4y)^9:9!$ as for $s = 4$, $a = n\ (s-1) = 9$. The factor K is found by replacing in the above polynomial $(3y)^i:i!$ by $\left(\frac{3}{4}\right)^i\binom{a+i}{i}$, as follows

$$K = \sum_0^6 \left(\frac{3}{4}\right)^i\binom{9+i}{i} + \frac{42}{43}\left(\frac{3}{4}\right)^7\binom{16}{7}$$
$$+ \frac{112}{129}\left(\frac{3}{4}\right)^8\binom{17}{8} + \frac{224}{387}\left(\frac{3}{4}\right)^9\binom{18}{9}.$$

The coefficients k_{a+1} et cetera of the polynomial of the denominator are determined by

$$k_{a+1} = \frac{K-1}{K}$$

$$k_{a+2} = \frac{K - 1 - \frac{3}{4}\binom{10}{1}}{K}$$

$$k_{a+3} = \frac{K - 1 - \frac{3}{4}\binom{10}{1} - \left(\frac{3}{4}\right)^2\binom{11}{2}}{K}$$

$$\cdots$$

6. *General Loss Equation for Complete 3-Split Homogeneous Gradings of the Second Order for* $y \rightarrow \infty$

The polynomial of the numerator consists, as already stated, of $n\binom{s-1}{2} + 1$ terms. Of these, $n\ (s-2) + 1$ have the following general appearance.

$$[(s-1)\ y]^i:i!$$

and $n\binom{s-2}{2}$ the appearance of

$$f_i[(s-1)\ y]^i:i!,$$

where f_i is a numerical factor. For three-split gradings, the latter group lapses, which fact permits the establishment of a general equation for three-split gradings.

The numerator contains first of all the factor

$$(sy)^a:a! = (3y)^{2n}:(2n)!$$

as for $s = 3$, $a = 2n$.

The numerator further includes the polynomial

$$\sum_0^n (2y)^i:i!.$$

The denominator includes the numerical factor

$$K = \sum_0^n \left(\frac{2}{3}\right)^i\binom{2n+i}{i}$$

and the polynomial

$$\sum_{0}^{2n} (3y)^i : i! + k_{2n+1}\,(3y)^{2n+1} : (2n+1)! + \cdots k_{3n}\,(3y)^{3n} : (3n)!$$

the coefficients of which are known by

$$k_{2n+1} = \sum_{1}^{n} \left(\frac{2}{3}\right)^i \binom{2n+i}{i} : K, \qquad k_{2n+2} = \sum_{2}^{n} \left(\frac{2}{3}\right)^i \binom{2n+1}{i} : K \cdots$$

The complete equation for the wanted grade of service for gradings with $s = 3$, therefore, is

$$W = \frac{\dfrac{(3y)^{2n}}{(2n)!} \displaystyle\sum_{0}^{n} \frac{(2y)^i}{i!}}{\displaystyle\sum_{0}^{n} \left(\frac{2}{3}\right)^i \binom{2n+i}{i} \times \sum_{0}^{2n} \frac{(3y)^i}{i!} + \sum_{j=1}^{j=n} \left[\frac{(3y)^{2n+j}}{(2n+j)!} \sum_{i=j}^{i=n} \left(\frac{2}{3}\right)^i \binom{2n+i}{i} \right]}. \tag{7}$$

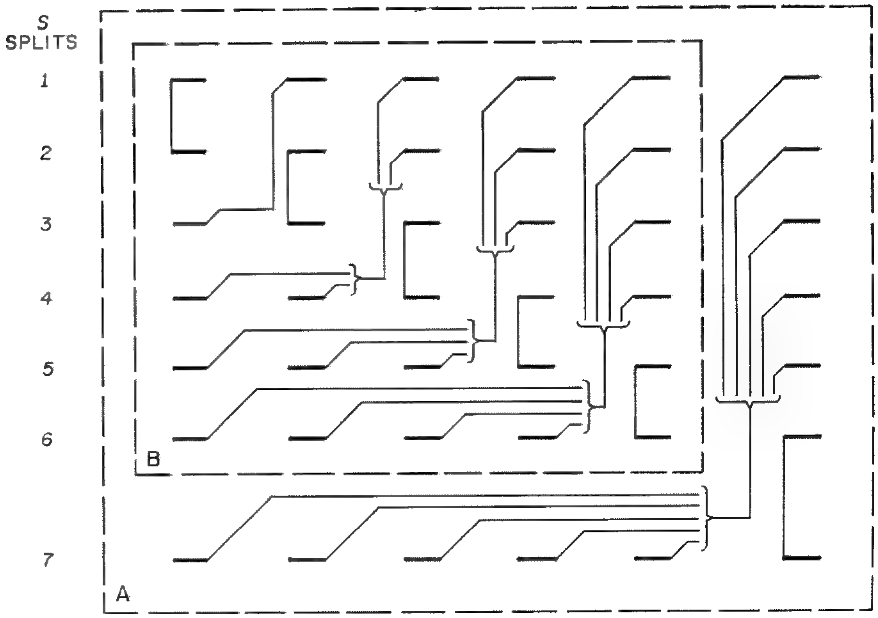


Figure 2—Relation between two consecutive gradings A and B. Only one complete cycle is shown. ($n = 1$.)

7. General Recurrent-Loss Equation ($y \rightarrow \infty$)

From the groups of equations appearing in section 3, the following general loss equation for homogeneous gradings of the second order having s splits and n outlets per subgroup valid for the limit $y \rightarrow \infty$, can readily be derived.

$$\boxed{W_{n,s}(sy) = \frac{{}_N P_{n,s}\,(sy)}{{}_{N-a}P_{n,s-1}\,[(s-1)y]},} \tag{9}$$

where ${}_N P_{n,s}\,(sy)$ represents the probability of finding all N outlets of the considered grading occupied, that is

$${}_N P_{n,s}\,(sy) = \frac{{}_N k_{n,s}\,(sy)^N : N!}{\displaystyle\sum_{0}^{a} (sy)^i : i! + \sum_{a+1}^{N} {}_i k_{n,s}\,(sy)^i : i!}. \tag{10}$$

The relation between two consecutive gradings, A and B in s is illustrated in Figure 2.

The same equation can also be written in a form more similar to the expressions used in the preceding sections.

$$\boxed{W_{n,s}\,(sy) = \frac{A_{n,s-1}[(s-1)y]}{A_{n,s}\,(sy)} \times \frac{(sy)^a : a!}{K_{n,s-1}},} \tag{11}$$

where

$$\sum_{0}^{a} (sy)^i : i! + \sum_{a+1}^{N} {}_i k_{n,s}\,(sy)^i : i! = A_{n,s}\,(sy). \tag{12}$$

$$\sum_{0}^{a-n}\left(\frac{s-1}{s}\right)^{i}\binom{a+i}{i}+\sum_{a-n+1}^{N-a}{}_{i}k_{n,s-1}\left(\frac{s-1}{s}\right)^{i}\binom{a+i}{i}=K_{n,s-1}. \tag{13}$$

$$\left.\begin{aligned}
\left[\sum_{1}^{a-n}\left(\frac{s-1}{s}\right)^{i}\binom{a+i}{i}+\sum_{a-n+1}^{N-a}{}_{i}k_{n,s-1}\left(\frac{s-1}{s}\right)^{i}\binom{a+i}{i}\right]\frac{1}{K_{n,s-1}}&={}_{a+1}k_{n,s}\\
\left[\sum_{2}^{a-n}\left(\frac{s-1}{s}\right)^{i}\binom{a+i}{i}+\sum_{a-n+1}^{N-a}{}_{i}k_{n,s-1}\left(\frac{s-1}{s}\right)^{i}\binom{a+i}{i}\right]\frac{1}{K_{n,s-1}}&={}_{a+2}k_{n,s}\\
\left[\sum_{a-n+1}^{N-a}{}_{i}k_{n,s-1}\left(\frac{s-1}{s}\right)^{i}\binom{a+i}{i}\right]\frac{1}{K_{n,s-1}}&={}_{2a-n+1}k_{n,s}\\
\left[{}_{N-a}k_{n,s-1}\left(\frac{s-1}{s}\right)^{N-a}\binom{N}{N-a}\right]\frac{1}{K_{n,s-1}}&={}_{N}k_{n,s}.
\end{aligned}\right\} \tag{14}$$

$$n=1,2,3\cdots \qquad s=3,4\cdots \qquad a=n(s-1) \qquad N=n\binom{s}{2}.$$

8. Traffic Emanating from Limited Number of Sources

By analogy with similar problems, it may be concluded that in case the traffic emanates from a limited number of sources the above equations are valid but that the probabilities of the Poisson type should be replaced by those of the Bernouilli type.

9. Traffic Tables

The computation of the data contained in the following tables was performed in the electronic computer developed by the Bell Telephone Manufacturing Company for the Institut de Recherche Scientifique pour l'Industrie et l'Agriculture and for the Fonds National de la Recherche Scientifique, two Belgian research centers. It is operated by the Centre d'Etude et d'Exploitation des Calculateurs Electroniques and this program was prepared by Miss M. Lietaert.

The values shown in the tables represent y, that is, the traffic offered to each split of a number of homogeneous gradings expressed in erlangs. The results are given in floating-decimal form, the last digit of each entry being the power of 10 by which the mantissa has to be multiplied.

$n = 1$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.819	−02	0.261	−01	0.372	−01	0.596	−01	0.856	−01	0.124	−00	0.208	−00
4	0.531	−01	0.117	−00	0.149	−00	0.206	−00	0.266	−00	0.347	−00	0.505	−00
5	0.148	−00	0.273	−00	0.329	−00	0.425	−00	0.519	−00	0.640	−00	0.867	−00
6	0.292	−00	0.481	−00	0.562	−00	0.695	−00	0.821	−00	0.980	−00	0.126	01
7	0.476	−00	0.730	−00	0.834	−00	0.100	01	0.115	01	0.135	01	0.169	01
8	0.695	−00	0.101	01	0.113	01	0.133	01	0.152	01	0.174	01	0.214	01
9	0.944	−00	0.132	01	0.146	01	0.169	01	0.190	01	0.215	01	0.260	01
10	0.121	01	0.165	01	0.181	01	0.207	01	0.230	01	0.258	01	0.306	01
11	0.151	01	0.199	01	0.218	01	0.246	01	0.271	01	0.302	01	0.354	01
12	0.182	01	0.235	01	0.256	01	0.286	01	0.313	01	0.346	01	0.403	01

$n = 2$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.139	−00	0.255	−00	0.308	−00	0.399	−00	0.489	−00	0.605	−00	0.828	−00
4	0.455	−00	0.697	−00	0.798	−00	0.959	−00	0.111	01	0.130	01	0.164	01
5	0.911	−00	0.127	01	0.141	01	0.163	01	0.184	01	0.209	01	0.253	01
6	0.146	01	0.193	01	0.211	01	0.239	01	0.264	01	0.294	01	0.347	01
7	0.209	01	0.266	01	0.287	01	0.320	01	0.348	01	0.383	01	0.445	01
8	0.277	01	0.343	01	0.367	01	0.404	01	0.436	01	0.475	01	0.544	01
9	0.350	01	0.424	01	0.450	01	0.491	01	0.526	01	0.569	01	0.644	01
10	0.428	01	0.509	01	0.536	01	0.580	01	0.618	01	0.664	01	0.745	01
11	0.511	01	0.598	01	0.625	01	0.671	01	0.712	01	0.760	01	0.846	01

$n = 3$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.433	−00	0.665	−00	0.760	−00	0.917	−00	0.106	01	0.125	01	0.159	01
4	0.113	01	0.154	01	0.170	01	0.195	01	0.217	01	0.245	01	0.294	01
5	0.204	01	0.259	01	0.280	01	0.312	01	0.341	01	0.376	01	0.438	01
6	0.306	01	0.375	01	0.400	01	0.438	01	0.472	01	0.514	01	0.587	01
7	0.418	01	0.498	01	0.526	01	0.571	01	0.609	01	0.655	01	0.737	01
8	0.533	01	0.626	01	0.657	01	0.705	01	0.748	01	0.799	01	0.891	01
9	0.659	01	0.757	01	0.793	01	0.844	01	0.890	01	0.945	01	0.104	02
10	0.780	01	0.890	01	0.932	01	0.985	01	0.103	02	0.109	02	0.119	02
11	0.904	01	0.102	02	0.107	02	0.112	02	0.114	02	0.124	02	0.134	02

$n = 4$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.842	−00	0.118	01	0.131	01	0.153	01	0.173	01	0.198	01	0.243	01
4	0.198	01	0.253	01	0.273	01	0.305	01	0.334	01	0.369	01	0.432	01
5	0.336	01	0.407	01	0.433	01	0.473	01	0.509	01	0.553	01	0.630	01
6	0.488	01	0.573	01	0.604	01	0.650	01	0.692	01	0.743	01	0.833	01
7	0.649	01	0.747	01	0.781	01	0.833	01	0.880	01	0.936	01	0.103	02
8	0.818	01	0.925	01	0.963	01	0.102	02	0.107	02	0.113	02	0.124	02
9	0.994	01	0.110	02	0.114	02	0.120	02	0.126	02	0.133	02	0.145	02
10	0.120	02	0.128	02	0.133	02	0.138	02	0.144	02	0.153	02	0.164	02

$n = 5$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.133	01	0.177	01	0.194	01	0.221	01	0.245	01	0.276	01	0.331	01
4	0.293	01	0.359	01	0.384	01	0.422	01	0.457	01	0.499	01	0.574	01
5	0.479	01	0.564	01	0.595	01	0.642	01	0.684	01	0.735	01	0.826	01
6	0.681	01	0.780	01	0.816	01	0.870	01	0.918	01	0.977	01	0.108	02
7	0.894	01	0.100	02	0.104	02	0.110	02	0.115	02	0.122	02	0.134	02
8	0.111	02	0.122	02	0.126	02	0.133	02	0.138	02	0.146	02	0.160	02
9	0.133	02	0.145	02	0.149	02	0.156	02	0.163	02	0.172	02	0.190	02

$n = 6$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.187	01	0.240	01	0.261	01	0.292	01	0.321	01	0.357	01	0.421	01
4	0.395	01	0.472	01	0.500	01	0.544	01	0.584	01	0.632	01	0.718	01
5	0.629	01	0.727	01	0.761	01	0.815	01	0.862	01	0.921	01	0.102	02
6	0.883	01	0.993	01	0.103	02	0.109	02	0.114	02	0.121	02	0.133	02
7	0.115	02	0.127	02	0.130	02	0.137	02	0.144	02	0.151	02	0.165	02
8	0.143	02	0.155	02	0.159	02	0.166	02	0.175	02	0.181	02	0.197	02

$n = 7$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.246	01	0.307	01	0.331	01	0.367	01	0.400	01	0.441	01	0.514	01
4	0.502	01	0.589	01	0.621	01	0.670	01	0.714	01	0.768	01	0.865	01
5	0.787	01	0.894	01	0.932	01	0.991	01	0.104	02	0.110	02	0.122	02
6	0.109	02	0.121	02	0.125	02	0.132	02	0.138	02	0.145	02	0.159	02
7	0.141	02	0.155	02	0.159	02	0.166	02	0.173	02	0.181	02	0.197	02
8	0.176	02	0.188	02	0.194	02	0.202	02	0.212	02	0.218	02	0.236	02

$n = 8$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.308	01	0.378	01	0.404	01	0.444	01	0.482	01	0.526	01	0.608	01
4	0.612	01	0.708	01	0.744	01	0.798	01	0.847	01	0.906	01	0.101	02
5	0.948	01	0.106	02	0.110	02	0.117	02	0.122	02	0.129	02	0.142	02
6	0.130	02	0.143	02	0.147	02	0.155	02	0.161	02	0.169	02	0.184	02
7	0.169	02	0.182	02	0.186	02	0.195	02	0.203	02	0.211	02	0.228	02
8	0.209	02	0.220	02	0.229	02	0.238	02	0.249	02	0.254	02	0.274	02

$n = 9$

s \ W	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.373	01	0.450	01	0.479	01	0.523	01	0.564	01	0.614	01	0.703	01
4	0.727	01	0.830	01	0.868	01	0.927	01	0.980	01	0.104	02	0.116	02
5	0.111	02	0.123	02	0.128	02	0.135	02	0.141	02	0.148	02	0.163	02
6	0.152	02	0.165	02	0.170	02	0.179	02	0.185	02	0.193	02	0.210	02

$n = 10$

s \ W	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.440	01	0.524	01	0.555	01	0.604	01	0.648	01	0.702	01	0.799	01
4	0.843	01	0.955	01	0.996	01	0.105	02	0.111	02	0.118	02	0.131	02
5	0.127	02	0.141	02	0.145	02	0.153	02	0.160	02	0.168	02	0.183	02
6	0.174	02	0.188	02	0.193	02	0.203	02	0.210	02	0.218	02	0.236	02

$n = 11$

s \ W	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.509	01	0.601	01	0.633	01	0.686	01	0.733	01	0.792	01	0.896	01
4	0.961	01	0.108	02	0.112	02	0.119	02	0.125	02	0.132	02	0.146	02
5	0.144	02	0.158	02	0.163	02	0.171	02	0.178	02	0.187	02	0.203	02
6	0.195	02	0.210	02	0.216	02	0.225	02	0.233	02	0.244	02	0.261	02

$n = 12$

s \ W	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.580	01	0.677	01	0.713	01	0.769	01	0.820	01	0.882	01	0.993	01
4	0.108	02	0.120	02	0.125	02	0.132	02	0.139	02	0.146	02	0.161	02
5	0.161	02	0.175	02	0.181	02	0.189	02	0.197	02	0.206	02	0.223	02

$n = 13$

s \ W	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.650	01	0.755	01	0.794	01	0.853	01	0.907	01	0.973	01	0.109	02
4	0.120	02	0.133	02	0.138	02	0.145	02	0.152	02	0.161	02	0.176	02
5	0.178	02	0.192	02	0.199	02	0.207	02	0.215	02	0.225	02	0.245	02

$n = 14$

$s \backslash W$	0.0001		0.001		0.002		0.005		0.01		0.02		0.05	
3	0.725	01	0.834	01	0.875	01	0.937	01	0.994	01	0.106	02	0.119	02
4	0.133	02	0.146	02	0.151	02	0.160	02	0.167	02	0.174	02	0.191	02
5	0.195	02	0.209	02	0.217	02	0.225	02	0.234	02	0.244	02	0.266	02

United States Patents Issued to International Telephone and Telegraph System; August–October 1957

Between August 1 and October 31, 1957, the United States Patent Office issued 49 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

A. H. W. Beck, Standard Telecommunication Laboratories (London), Electron Discharge Apparatus, 2 810 853.

W. Berthold, C. Lorenz A. G. (Stuttgart), Gun System for Cathode-Ray Tubes, 2 802 139.

E. M. Bradburd, Federal Telecommunication Laboratories, Tuned High-Frequency Amplifier, 2 803 710.

A. E. Brewster, Standard Telecommunication Laboratories (London), Regenerative Telegraph Repeaters, 2 802 052.

J. H. Bryant, Federal Telecommunication Laboratories, Radio-Frequency Matching Devices, 2 803 777.

H. Burr, Standard Telephones and Cables (London), Electric Cables, 2 810 011.

K. W. Cattermole, Standard Telecommunication Laboratories (London), Electric Pulse Generators Employing Semiconductors, 2 807 719.

D. W. Davis, Farnsworth Electronics Company, Cathode-Ray Amplifier, 2 808 526.

R. C. Davis and E. B. Moore, Federal Telephone and Radio Company, Combination Automatic-Gain-Control and Silencer Amplifier, 2 802 099.

C. L. Day, Capehart-Farnsworth Company, Vacuum-Tube Element, 2 802 126.

E. de Faymoreau, Federal Telecommunication Laboratories, Servomotor Control System, 2 810 874.

M. J. Di Toro, W. Graham, and S. M. Schreiner, Federal Telecommunication Laboratories, Compressed-Frequency Communication System, 2 810 787.

E. L. Earle, Kellogg Switchboard and Supply Company, Armature Keeper for Electromagnetic Relay, 2 811 681.

H. F. Engelmann, Federal Telecommunication Laboratories, Attenuators, 2 810 891.

P. F. M. Gloess, Le Matériel Téléphonique (Paris), Sounding Device Using Electromagnetic Waves, 2 807 016.

F. P. Gohorel, Compagnie Générale de Constructions Téléphoniques (Paris), Automatic Telephone Systems, 2 810 018.

A. N. Gulnick, Federal Telecommunication Laboratories, Delayed Action Switch, 2 810 797.

T. F. S. Hargreaves, H. T. Prior, and W. F. S. Chittleburgh, Standard Telephones and Cables (London), Voice-Frequency-Signal Receivers, 2 806 903.

E. J. Hasney, Capehart–Farnsworth Company, Shaft Coupling Assembly, 2 801 531.

A. Q. Hislop and W. Q. Leysath, Capehart–Farnsworth Company, Stroboscopic Device, 2 802 145.

L. Holik and H. Nowotny, Vereinigte Telephon- und Telegraphenfabriks, A. G., Czeija, Nissl & Co. (Vienna), Electrolytic Capacitors, 2 806 982.

J. F. Houdek, Jr., Kellogg Switchboard and Supply Company, Selective Gong Damper for Dual-Gong Ringer, 2 808 019.

T. M. Jackson and A. D. Odell, Standard Telephones and Cables (London), Electrical Circuits Using Multigap Cold-Cathode Gas-Filled Tubes, 2 810 861.

A. G. Kandoian, Federal Telephone and Radio Company, Antenna Couplings, 2 807 713.

W. Keilig, Mix & Genest (Stuttgart), Edgewise Conveyor System, 2 809 741.

M. Kenmoku, Nippon Electric Company (Tokyo), Traveling-Wave Tube, 2 807 742.

E. Labin and D. D. Grieg, Federal Telephone and Radio Company, Indicating and Calibrating Means, 2 802 179.

G. X. Lens, Bell Telephone Manufacturing Company (Antwerp), Combined Code Recorder and Selector, 2 807 376.

A. Lesti, Federal Telecommunication Laboratories, Decoder for Pulse-Code-Modulation Systems, 2 807 715.

A. Lieb, C. Lorenz A. G. (Stuttgart), Electronic Controlling Device, 2 807 738.

F. Malsch, C. Lorenz A. G. (Stuttgart), Tuning Indicator Valve, 2 802 128.

A. W. McEwan, C. E. Anderson, and C. V. Daniels, Federal Telecommunication Laboratories, Traveling-Wave Electron Discharge Device, 2 806 973.

E. M. S. McWhiter, Intelex Systems Incorporated, Document Jacketing and Encoding Machine, 2 806 335.

R. B. Odden, Capehart–Farnsworth Company, Wave-Selecting and Synchronizing System, 2 802 105.

S. B. Pickles and M. A. Karpeles, Federal Telecommunication Laboratories, Radio Navigation Receiver, 2 803 821.

W. H. P. Pouliart and J. P. H. Vandevenne, Bell Telephone Manufacturing Company (Antwerp), Electrical Information Storage Arrangement, 2 807 004.

H. T. Prior and W. F. S. Chittleburgh, Standard Telephones and Cables (London), Synchronizing Arrangements in Electric Telegraph Systems, 2 802 051.

F. Schmidt, Mix & Genest (Stuttgart), Circuit Arrangement to Improve Impulsing and Supply in Stations with Long Subscribers' Lines, 2 806 085.

W. Sindzinski, Mix & Genest (Stuttgart), Unloader for Edgewise Conveyor System, 2 810 469.

W. Sindzinski, Mix & Genest (Stuttgart), Coacting Belt Conveyor System, 2 806 582.

W. Sindzinski and M. Muller, Mix & Genest (Stuttgart), Conveying Device, 2 801 725.

A. T. Starr, H. Grayson, R. A. G. Dunkley, and T. H. Walker, Standard Telecommunication Laboratories (London), Electromagnetic Switches, 2 802 170.

E. Stein and F. H. Numrich, Federal Telecommunication Laboratories, Protective Circuit, 2 810 858.

K. Steinbuch, Mix & Genest (Stuttgart), Circuit Arrangement to Change Characteristic Curve of Multielectrode Tubes, 2 806 154.

H. K. J. Strosche, Süddeutsche Apparatefabrik (Nürnberg), Method of Producing Selenium Rectifiers, 2 807 762.

I. R. Taylor, Federal Telecommunication Laboratories, Antenna Structure, 2 805 414.

F. A. Termini and J. M. Rafalko, Federal Telecommunication Laboratories, Apparatus for Printed-Circuit Solder Coating, 2 803 216.

S. H. Towner and L. R. Hatch, Standard Telephones and Cables (London), Electromagnetic Light-Current Contact-Making Relays, 2 802 156.

T. H. Walker, Standard Telecommunication Laboratories (London), Electric Trigger Circuits, 2 806 153.

Sounding Device Using Electromagnetic Waves

2 807 016

P. F. M. Gloess

This basic patent covers plan-position-indicator radar. The invention covers broadly the arrangement for deriving directly from reflected radar pulses information including the angular position of the reflecting object and the distance to the object, this combined information being displayed on a single indicator.

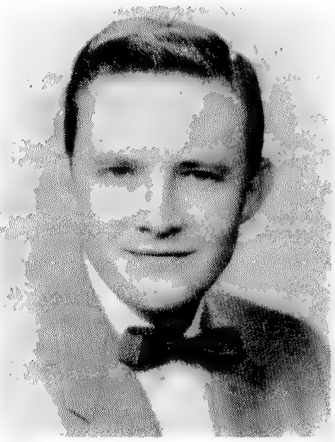
Document Jacketing and Encoding Machine

2 806 335

E. M. S. McWhirter

This invention covers the conveyor and jacketing machine for use in automatic bank accounting machines.

Contributors in This Issue



Guy E. Adams

Guy E. Adams holds the B.S.E.E. and B.S.M.E. degrees from Purdue University and has attended its graduate school. He has had extensive experience in control systems for drone aircraft. During the past several years he has been engaged in the analysis, design, and development of control systems and computers for missile control, fire control, bombing systems, and simulators. As a senior engineer in the electromechanical section at Farnsworth Electronics Company, he prepared the paper in this issue on servomechanisms.

At the present time, Mr. Adams is systems section head in the inertial systems department, mechanical division, of General Mills, Incorporated. He is a Senior Member of the IRE.

F. Buchholtz

• • •

F. Buchholtz was born in Braunschweig in Germany on June 4, 1923. His engineering education was received at the State College in Dortmund.

After working as a development engineer for various organizations, he joined Mix & Genest, now a division of Standard Elektrik A.G., in 1951.

His particular responsibility is the development of amplifiers and repeaters for carrier-operated telephone systems and he is head of that laboratory within Standard Central Laboratories in Stuttgart. Mr. Buchholtz reports on the *Z6NC* carrier telephone system in this issue.

• • •

Leopold Christiansen

Leopold Christiansen was born in Kiev, Russia, on December 2, 1913. He studied at the Technical University of Stuttgart, at the Rensselaer Polytechnic Institute of Troy, New York, and at the Technical University of Dresden, where he received his diploma and his doctorate.

He worked first at the Institute of Electrical Engineering of the Technical University of Dresden. In 1948, he joined Mix & Genest, now a division of Standard Elektrik A.G. Since that time, he has been engaged in the development of various carrier-frequency telephony systems, and is now head of the wire and cable transmission laboratories.

In this issue, he is coauthor of the paper on the *Z6NC* carrier-frequency telephone system.

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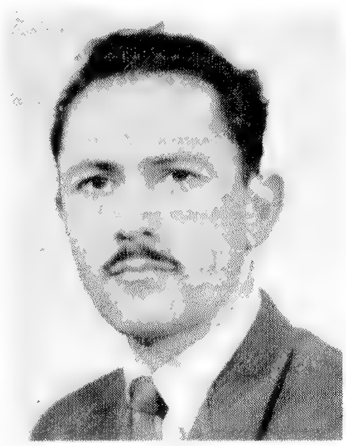


Lucien R. Gillon

Lucien R. Gillon was born in Saint-Maur-des-Fossés, France, on February 21, 1920. He graduated in 1940 as an engineer from the Ecole Nationale d'Arts et Métiers of Paris.

He joined the Compagnie Générale de Constructions Téléphoniques in February 1941 and was assigned to the remote-control division of the technical department. Mr. Gillon has been the head of this division since 1953. He is the author of the paper in this issue on the supervision equipment for a 65-kilovolt power network.

• • •

Jakob Kruithof was born in Rotterdam, Netherlands, on November 6, 1894. He received the degree of Electro-Technical Engineer from Delft Technical High School in Holland in 1922 and of Doctor in Applied Sciences from Ghent University in Belgium in 1945.

Dr. Kruithof is Technical Development Administrator of Bell Telephone

Jakob Kruithof

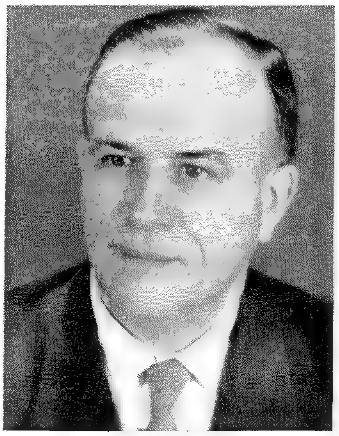
Manufacturing Company in Antwerp, Belgium, and Assistant Technical Director of the International Standard Electric Corporation. He reports in this issue on homogeneous gradings for telephone switching systems.

• • •

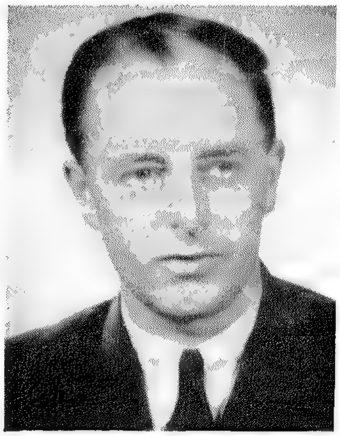
Wilhelm Schmitz was born in Honnef on the Rhine in Germany on April 14, 1902. He received a diploma from the Technical College of Aachen in 1926, a doctor of engineering degree from the Technical College of Berlin in 1932, and, from the same college, a doctor of engineering license in 1939.

After working for Siemens and Halske from 1927 to 1949, he joined C. Lorenz A.G. and is now director of the laboratory for railroad safety systems.

Dr. Schmitz holds a large number of patents and has published his work extensively. His paper in this issue describes the Tricon railroad interlock system.



Wilhelm Schmitz



W. Zaiser

W. Zaiser was born on July 27, 1912 in Boeblingen, Germany. He received a doctorate from the Technical Colleg of Stuttgart.

After several years with Allgemein Elektrizitäts Gesellschaft, he joine Mix and Genest, now a division o Standard Elektrik A.G., in 1951. He i now in charge of the laboratory for th development of multichannel telephon carrier systems of Standard Centra Laboratories in Stuttgart. Dr. Zaiser i coauthor of the paper on the *Z6N* carrier system.

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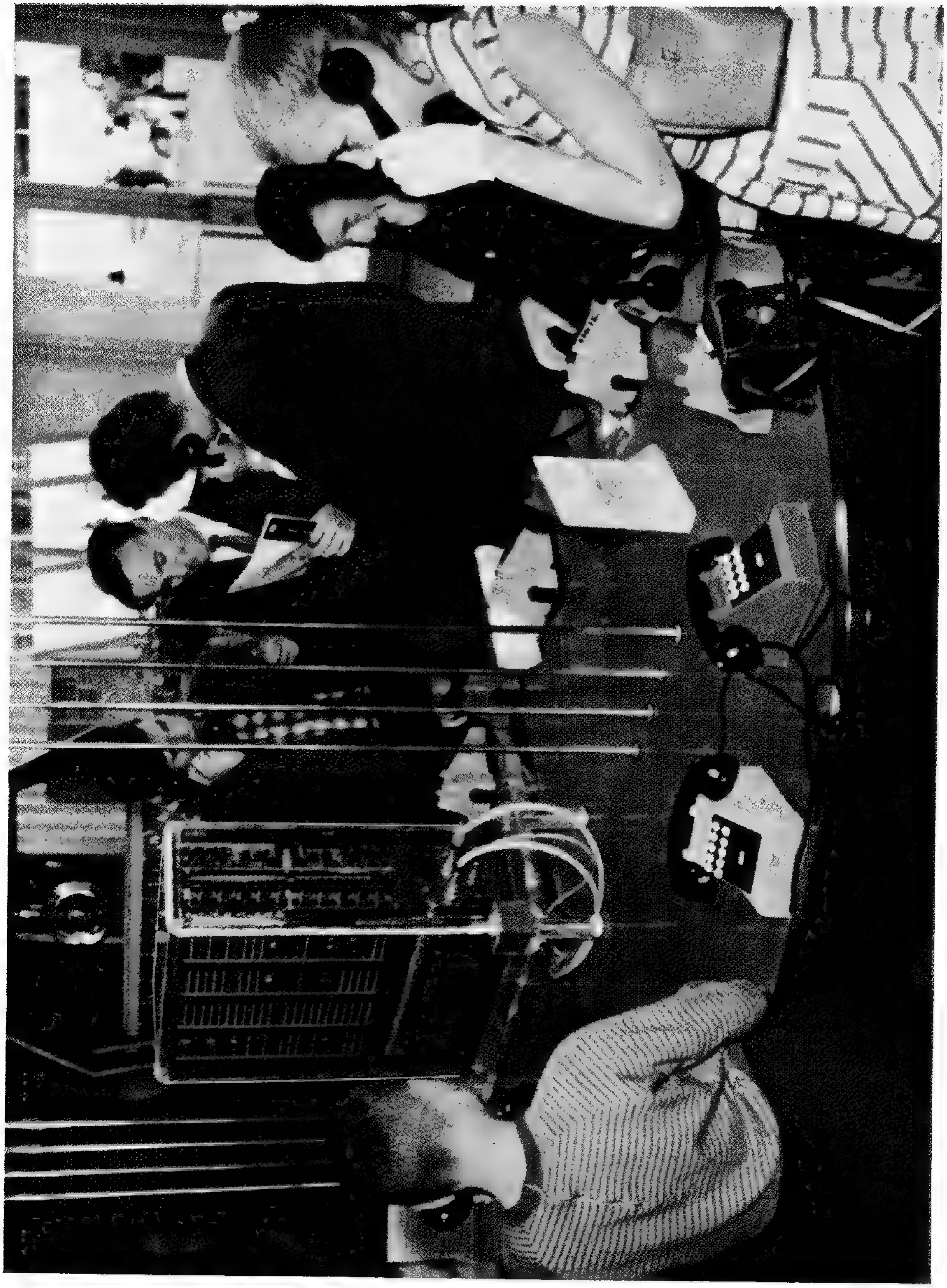


Electronic 20-Line Private Automatic Telephone Exchange

DEVELOPED and manufactured by Laboratoire Central de Télécommunications, Paris, and by Bell Telephone Manufacturing Company, Antwerp, a fully electronic 20-line private automatic branch exchange was a center of interest at Bell's pavilion at the 1958 Brussels World Fair.

The frontispiece photograph shows typical visitors from the stream that continually placed call after call to each other using the telephones on the table and marvelled to see their connections quickly established by purely electronic means.

The exchange can handle 4 simultaneous conversations and will process 2 calls simultaneously by means of 20 subscribers' line circuits, 4 junction circuits, and 2 registers. The fundamental circuit elements used[1] are silicon junction speech-switching diodes controlled by magnetic flip-flop circuits made of a saturable reactor in series with a capacitor to form a ferroresonant circuit. By completely excluding moving contacts, such as those on relays, the life of the equipment becomes practically indefinite.

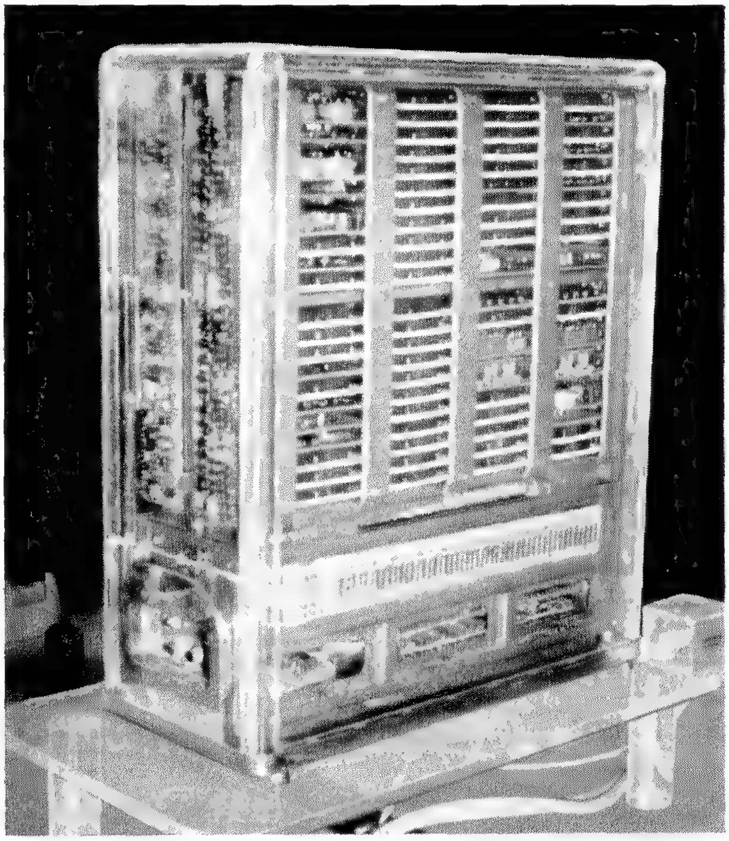
The subscriber's set used with the equipment differs from the conventional design in that digits are transmitted from a keyboard and not from a dial and that the bell has been replaced by an electroacoustical device driven by a transistor amplifier in the set.

In the exchange, the subscriber's line circuit incorporates a line transformer and the device setting the condition of the line, busy or free, et cetera. The silicon-diode speech-switching circuits use two miniature diodes to establish a speech path between subscribers. In the blocking condition, the diodes are equivalent to a 1000-megohm resistor in parallel with a 5 picofarad capacitor. In the conducting condition, they have a resistance of only 4 ohms.

The peak power transmitted is 50 milliwatts and the total attenuation from line to line is 1 decibel.

The diode switches are operated by the magnetic flip-flops, which also form the register counters. These flip-flops are driven by an 8-kilocycle-per-second 10-volt supply. At this voltage, they have two operating conditions; in one state, the current passed is 15 times that of the other condition. After conversion to direct current by selenium rectifiers, the flip-flop output polarizes the diodes in conducting or blocking conditions.

Printed-circuit construction is used in the exchange; its dimensions are but 22 by 53 by 61 centimeters (8.8 by 21 by 24 inches). The power consumed by the exchange itself (excluding the microphone currents of the subscribers' sets) is only 30 watts at 24 volts.



[1] C. Dumousseau, "Fully Electronic 20-Line Automatic Telephone Exchange," *Electrical Communication,* volume 34, pages 92–101; June, 1957.

Electron-Beam Voltage-Indicator Tube EM84

By A. LIEB

Lorenz Werke, division of Standard Elektrik Lorenz A.G.; Stuttgart, Germany

INDICATOR TUBES are simplified cathode-ray tubes in which the magnitude of a voltage is made visible on a target by electron-optical means. The vacuum-tight envelope and the constructional principles correspond to those of ordinary vacuum tubes. The voltage indication of such tubes is not impaired by inertia effects as in the case of mechanical instruments and the tube is insensitive to reasonable voltage overloads. Moreover, the tube is usually so designed that no power is drawn from the measured circuit.

Commonly known indicator tubes have also been called magic eye or magic fan, depending on the shape of the luminous image. They are widely used in superheterodyne receivers with automatic gain control, where they facilitate tuning. In such applications, the tube indicates the magnitude of the gain-control voltage. Accurate tuning results in the maximum value of this voltage. Tuning by ear alone is difficult, as the automatic gain control causes almost the same volume level whether the receiver is tuned or detuned. Proper tuning is particularly important for sets with high selectivity and good reproduction as a slight detuning from the center of the carrier frequency will cause distortion; particularly with frequency-modulation receivers. Tuning-indicator tubes have therefore found wide application, particularly in Europe where the customer demands very-high-quality sound reproduction. Thus, 80 percent of all radio receivers made in Germany are equipped with tuning-indicator tubes.

In the United States, few receivers had such indicator tubes previously. However, the trend toward high-fidelity reproduction has brought about their wider use.

The principle of this tube dates back to the magic eye developed in the 1930's. The target of the magic eye is a metal funnel coated inside with luminous phosphor. An axial cathode with a parallel control electrode are inside the target. The control electrode produces a shadow on the target. The shadow varies in size depending on the voltage applied to the control electrode.

This magic-eye indicator tube has the disadvantage of a small target area. The concave target is mounted in the narrow end of the glass tube envelope. The person tuning a receiver often has to change position to observe the tube indication. Moreover, the target brightness of the tube has a relatively short life.

The observation angle of the *EM84* tube is much greater and the decrease of the luminous intensity during life is much smaller. The luminous image is simple and facilitates accurate reading. In addition, voltage values can be calibrated on a scale attached to the outside of the tube. This property of the tube makes it suitable for indicating not only minimum and maximum values of a voltage, as is the case with conventional tubes, but also specific voltages. Because of this expanded application, the tube is designated a voltage-indicator tube rather than a tuning-indicator tube.

1. Constructional Principles

The luminous pattern of an indicator tube can be observed best and from the greatest angle if this pattern is produced directly on the glass envelope. This is particularly true when the moving edge of the luminous image is on the envelope portion having the greatest curvature. These requirements are most conveniently met with a luminous pattern whose moving edge is at a right angle to the tube axis for any possible value indicated. The pattern of the new tube consists of two luminous bands parallel to the tube axis so that the length of both bands changes along the tube axis.

Figure 1A is a photograph of the tube and Figure 1B shows the luminous pattern. Distance A between the edges is a measure of the voltage. If the tube is placed in a radio set so that a portion of the side of the glass envelope protrudes from the panel, good observation is ensured from the whole area in front of the cabinet.

The principle described allows for simple calibration that can be accomplished, for instance, by projecting a thermometer-like scale against

the exterior wall of the glass bulb. Since the luminous pattern is directly on the bulb wall, all parallax errors that might affect the reading are precluded. The measuring scale can be calibrated after completion of the tube, thus compensating for manufacturing tolerances.

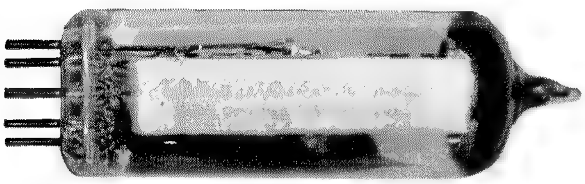


Figure 1A—Photograph of tube.

In conventional indicator tubes, zinc silicate is generally used for the target phosphor. This phosphor yields a relatively high light output in the green spectral range even if the excitation energy is small. However, the durability of this phosphor when exposed to electron bombardment is not very favorable. Beside zinc silicate, a suitably prepared phosphor of zinc oxide will also emit sufficient light and the decrease caused by electron bombardment is much less than for zinc silicate. This has been known for many years, but the bluish-green tint of the zinc-oxide emission did not favor its general introduction. Contrariwise, the zinc silicate produces an emission of pure green color. The pale color of the zinc oxide is caused by a high white component in the emitted light, which is less easily identifiable in ambient illumination than light from zinc silicate.

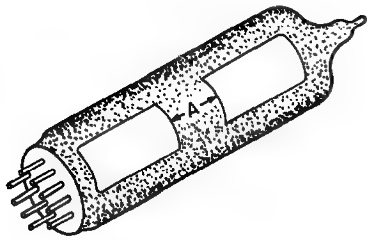


Figure 1B—Drawing of tube showing the fluorescent bands. Dimension *A* varies with the control voltage.

In the *EM84* indicator tube, however, a phosphor of zinc oxide is used. The problem of the emitted color has been solved by a color filter mounted so that it covers the luminous areas of the tube. The interfering white component thus serves a useful purpose; in addition, the advantage is obtained that the resulting color can be controlled by the spectral transmission factor of the filter. This advantage can be used to mark in a simple and effective way that portion of the luminous band that corresponds to a given voltage.

In the case under discussion, provision was made to change the color of the luminous bands as they expand along the scale. This arrangement will be seen in Figure 2; the tube bulb carries optical filters with different transmission factors at sections designated *F1* and *F2*. As soon as the luminous-band edges move over the distance *B*, the color of the bands will also change. The filter also increases the contrast and this is an advantage in bright daylight or artificial illumination.

The measured voltage is preamplified as in the case of conventional tubes. This ensures high indicating sensitivity and high input impedance. For economic reasons, the amplifier system is built into the indicator tube. The amplifier system of the indicator tube should be as simple as possible so that the tube can be manufactured on an economical basis and will give stable operation. In the new magic-band indicator tube, the design is simplified inasmuch as one cathode is utilized for both the indicator system and the amplifier system. The amplifier system is further simplified by using two rods to take the place of the usual control grid. These two control electrodes are arranged in the space-charge region of a grid at cathode potential. This grid surrounds the cathode and also functions as a grid for the indicator system.

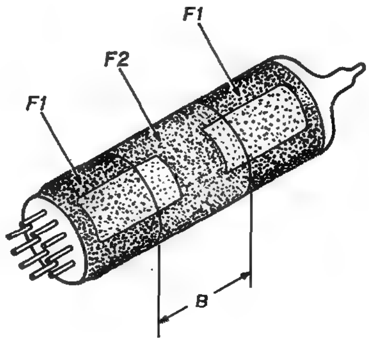


Figure 2—Colored filters can be placed over the fluorescent strip to identify certain values of measured voltage.

2. *Design Details*

Constructional details of the magic-band tube may be seen in Figure 3. The electrodes are held in place by the aperture frame, which is at target potential. This frame is welded to the leads passing through the glass button stem and has a slot-like aperture opposite the target on the glass bulb. The slot determines the width and the total length of the fluorescent patterns and, moreover, forms a shield between the mica spacers and the discharge space of the indicator system to prevent formation of disturbing charges on the mica spacers.

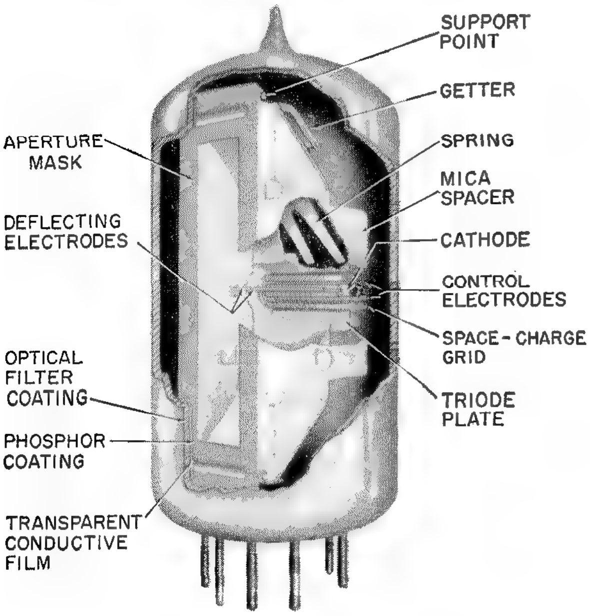


Figure 3—Electrode arrangement in the *EM84* voltage-indicator tube.

The target is on the inside wall of the bulb in front of the frame aperture. It consists of a transparent conductive film and the phosphor coating. The optical filter can be a transparent lacquer film applied to the exterior of the bulb. Instead of the lacquer, a colored transparent glass or heat-resistant plastic foil can be placed on the tube or in front of it.

By varying the transmission factor of the filter, a large number of various colors can be obtained. However, blue, green, and yellow have a particularly high light output owing to the preferred emission of the zinc-oxide phosphor.

For a two-color filter, it is expedient to select two complementary colors to achieve as great a color contrast as possible. The complementaries blue and yellow have the highest luminous intensity when zinc oxide is employed.

The length of the fluorescent bands, or the distance between their opposite ends, is varied by the deflecting electrodes. These deflecting electrodes are electrically connected to each other and this parallel connection results in a very-high deflection sensitivity and in particularly sharp edges of the fluorescent patterns.

The cathode is surrounded by a space-charge grid. That portion of the cathode facing the target serves as an electron source for the indicator system. The other side of the cathode forms the amplifier system. It consists of the control electrodes, the space-charge grid, and the plate. The control electrodes are two rods arranged between the cathode and space-charge grid. The amplifier section can be regarded as a special type of space-charge-controlled two-grid tube.

The electodes are supported by two mica spacers that, contrary to conventional design, are mounted parallel to the bulb axis. The electrode system is pressed against the glass bulb by the spring shown in Figure 3. Only a few points of the frame are in contact with the glass so that intolerable thermal glass strains during pumping are prevented. These points of support and the spring also ensure the parallel position of the

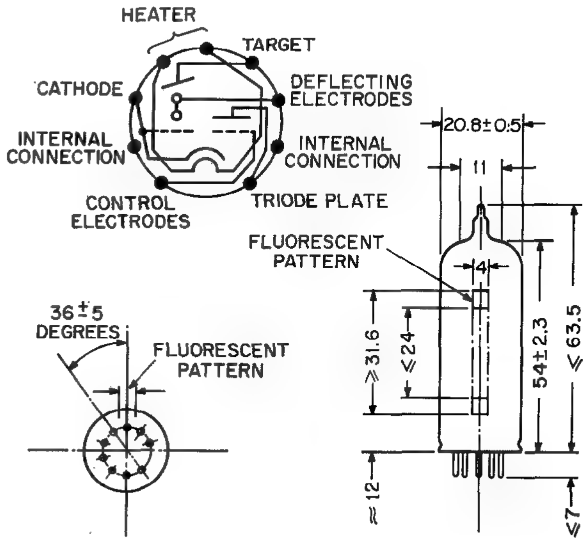


Figure 4—Socket connections, dimensions in millimeters, and position of fluorescent pattern on the tube.

frame with respect to the target. Finally, the points of support and the spring connect the target with the frame aperture mask so that both have the same potential.

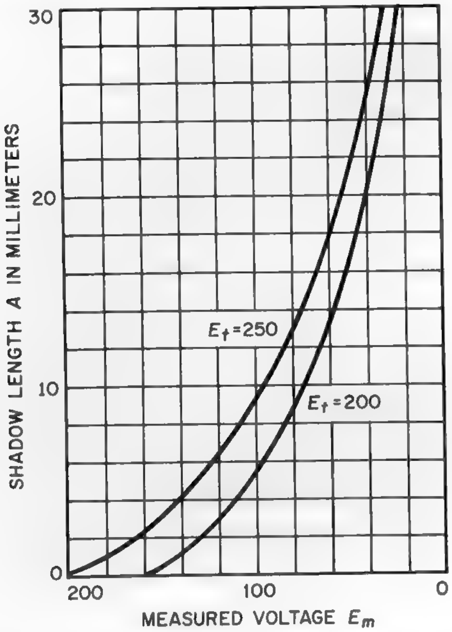


Figure 5—Shadow length A (see Figure 1B) as a function of the measured voltage E_m when applied directly to the deflecting electrodes. The target voltage = E_t.

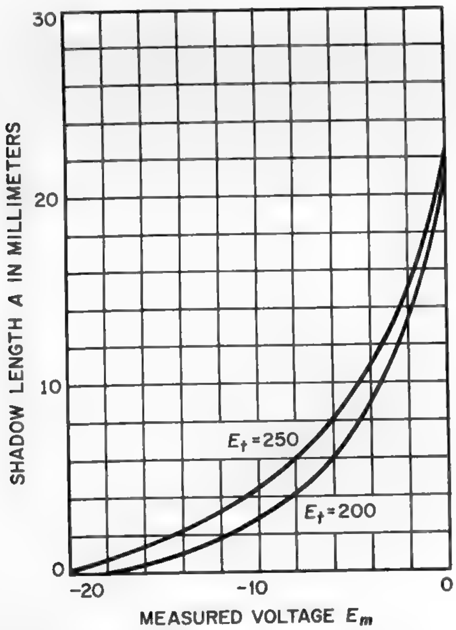


Figure 6—Shadow length A as a function of the measured voltage E_m when applied to the control electrode of the triode preamplifier. The target voltage = E_t. Triode plate load resistor = 470 kilohms.

3. *Properties of Tube*

The most-important properties of the *EM84* can be seen from Figure 4 and Table 1.

Figure 5 shows the distance A between the pattern edges as a function of the measured voltage when it is applied directly to the deflecting electrodes. Figure 6 is for that voltage applied to the amplifier control electrode. Figure 7 shows the derivative of A to the measured voltage depending on the measured voltage applied to the amplifier. This quantity to a first approximation is a measure of indicator sensitivity. A more-accurate evaluation of sensitivity should take into account not only the variation of A, but also its absolute value at any instant. When A becomes smaller, voltage variations are more-readily perceived, as the observer tends then to observe the doubly sensitive distance variations between the fluorescent edges rather than the changing length of one band.

TABLE 1

Operational Characteristics of *EM84*

Heater Values, Parallel Supply	
Heater Voltage	6.3 Volts ± 10 Percent
Heater Current	0.27 Ampere
Cathode	Oxide, Unipotential
Operational Values (Deflecting Electrodes Connected to Plate)	
Plate Voltage	250 Volts
Target Voltage	250 Volts
Plate Load Resistance	470 Kilohms
Control-Electrode Resistor	3.0 Megohms
Control-Electrode Bias	0 to −22 Volts
Plate Current	0.45 to 0.06 Milliampere
Target Current	1.1 to 1.6 Milliamperes
Shadow Length A	22 to 0 Millimeters
Maximum Values (Limiting)	
Cold Plate Voltage	550 Volts
Plate Voltage	300 Volts
Plate Dissipation	0.5 Watt
Cold Target Voltage	550 Volts
Target Voltage	300 Volts
Cathode Current	3.0 Milliamperes
Grid-Leak Resistance	3.0 Megohms
Heater-to-Cathode Voltage	100 Volts
Ambient Temperature Near Target	120 Degrees Centigrade

The subjective reading sensitivity was measured by plotting the smallest increment of voltage resulting in a perceptible change of the pattern over a range of measured voltages. These

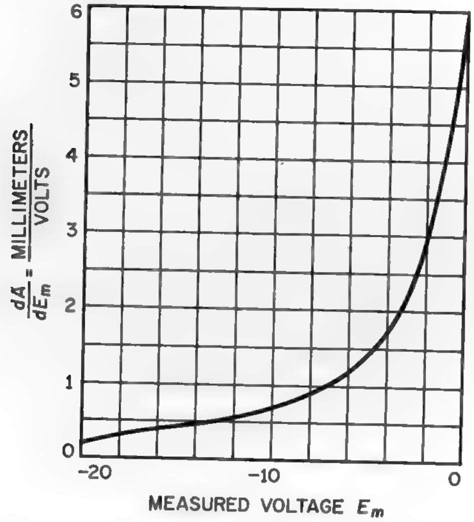


Figure 7—Derivative of length A to measured voltage E_m as a function of measured voltage applied to the grid of the triode preamplifier; Target voltage E_t = 250 volts.

observations were made at a practical distance of 40 centimeters by several persons independently. The resulting mean values obtained by all observers are shown in Figure 8. The measurements reveal the high sensitivity of the human eye to distance variations of parallel edges; perceptibility is far greater than for the varying arc of a luminous sector.

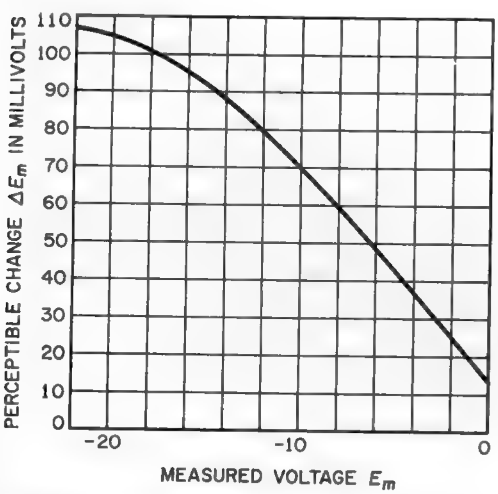


Figure 8—Smallest perceptible change ΔE_m in measured voltage E_m applied to control electrode.

Figure 9 is a plot of the target and plate currents as functions of the measured voltage applied to the amplifier system. The relatively low target current is particularly favorable for receivers with small power supplies.

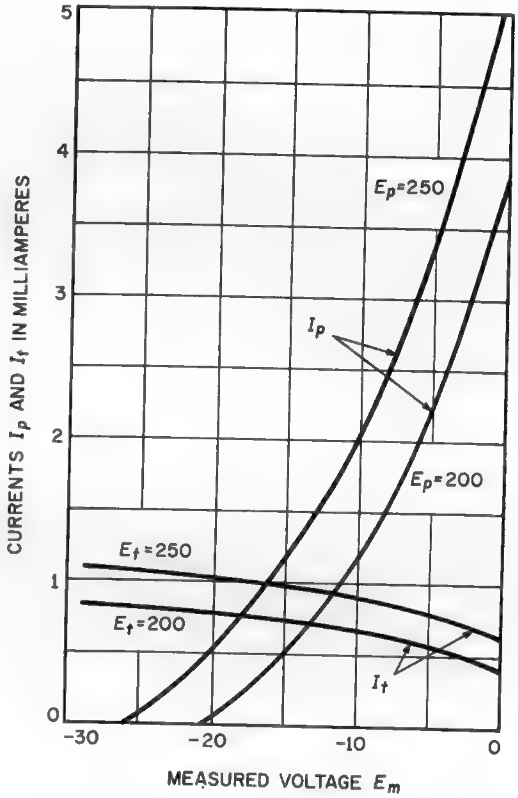


Figure 9—Target current I_t (with triode plate load resistance) and triode plate current I_p (without load resistance) as a function of measured voltage E_m applied to the control electrode. Target voltage = E_t and triode plate voltage = E_p.

4. *Applications*

The *EM84* tube, with the above advantages, can replace in all their applications the conventional magic-eye and magic-fan indicators.

Figure 10 shows the tube in a radio receiver. The tube is arranged above the tuning scale and in a horizontal position. The possibility of observing the tube from a wide angle is quite apparent with this arrangement. The tube is masked by a cylindrical bezel exposing only that portion of the tube bearing the target.

Another suitable type of mounting is immediately behind the dial or panel of a receiver. Since the target is on the glass envelope, the tube can be mounted so that the target appears in the plane of the tuning scale. In conventional tuning-indicator tubes, the luminous pattern is always set behind the scale so that observation is not as convenient as with the *EM84*.

The tube can even be moved along the scale with the scale pointer. The narrow fluorescent bands varying in length can be better adapted to the shape of the pointer than can the targets of conventional indicator tubes.

The possibility of marking certain values in a convenient scale on the exterior of the bulb, thus excluding parallax errors, makes the tube applicable to all cases where one or more voltage values must be determined. An example is the indication of overmodulation in a tape recorder.

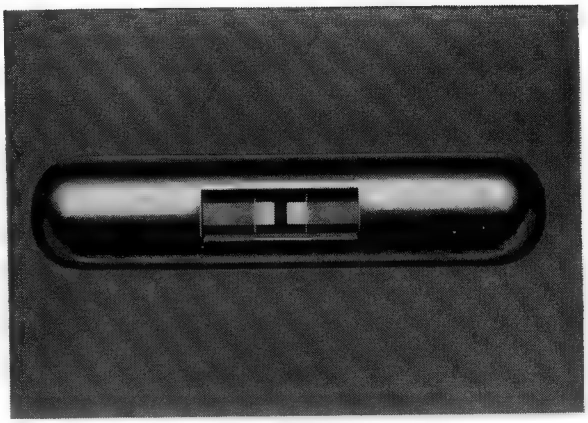


Figure 11—Application with colored filters as an overmodulation indicator in a tape recorder.

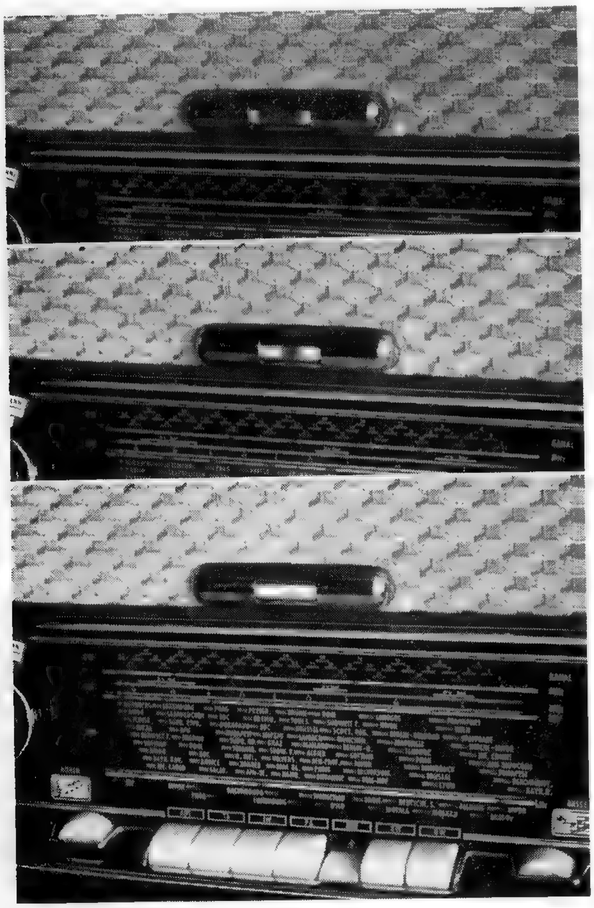


Figure 10—Type-*EM84* tube in receiver operating on weak (top), medium (middle), and strong (bottom) signals.

So far, tuning-indicator tubes used for this purpose have been arranged so that the fluorescent pattern moves visibly when the adjustment is correct, while maximum permissible amplitude results in disappearance of the shadow. This requires very-careful observation of the fluorescent pattern during the adjustment of the recorder gain and supervision from a distant place, as in home studios, is impossible. When the *EM84* is used in a tape recorder, two filters are employed as shown in Figure 11. The tube is covered by a mask so that only the target is visible. Voltages below the overload value are indicated in blue, while a yellow filter becomes effective as soon as overmodulation takes place.

By adding a third color filter, insufficient gain can be indicated in addition to the mean and overload values. In this case, the recorder must be adjusted so that the fluorescent pattern edges fluctuate only in the color area corresponding to permissible modulation. Adjustments of this kind can be easily comprehended and carried out even by unskilled personnel.

When a scale is fixed to the tube exterior, voltages can be measured with the *EM84*. Two measuring ranges are available without any parallel or series input resistors.

When the voltage to be measured is applied directly to the deflecting electrodes, the measuring range is +35-to-+200 volts. This range can be shifted to +20-to-+250 volts by changing the target voltage. For the case of the preamplified

voltage, the measuring range is 0-to-−22 volts. Here again the range can be shifted to −10-to-−30 volts by varying the plate supply voltage or the load resistance.

Figure 12 shows an *EM84* in such an application. The tube bulb carries a plastic-foil strip with a scale. To facilitate the reading, one of the fluorescent bands is completely masked. The sharp edge of the pattern permits very-accurate reading of the indicated values. The measuring accuracy is 5-to-15 percent due to variations in tube properties such as cathode emission, contact potential between cathode and control electrode, secondary-electron emission of the deflecting electrodes, and the like. Accuracy depends on the care taken in calibrating the scale and the measuring range selected. The tube will be of interest in certain fields of application where the main requirements on the measuring device are insensitivity to physical position, to vibration, and to overload; quick time response, low power drawn from the measured circuit, and economy.

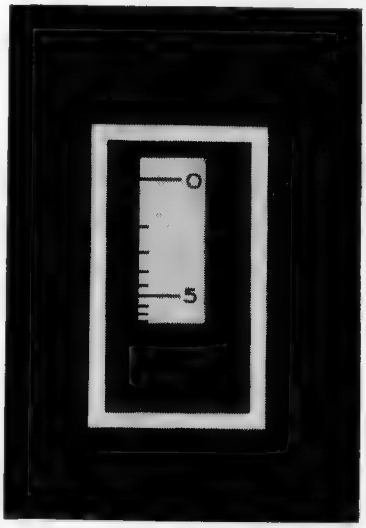


Figure 12—Tube used as a voltmeter.

Recent Telecommunication Development

Electronic Spectroanalysis

ELECTRONIC COMPUTERS have been adapted to the field of infrared spectroscopy to determine the elements that are present in an unknown chemical mixture. Each chemical element will absorb certain wavelengths of radiation to a known relative amount and the chief use of spectroscopy is to determine the elements in an unknown mixture from an examination of its pattern of absorbtivity.

If several elements are present in the specimen and their absorption lines are confusingly mixed together and even fall on the same wavelengths in some instances, the problem of sorting out which lines belong to which element is one that can take many hours or even days to solve by manual methods.

In cooperation with the Sloan–Kettering Institute for Cancer Research, the ITT Laboratories, a division of International Telephone and Telegraph Corporation, has produced equipment for the spectroscopic analysis of steroid compounds using the infrared spectrum. The spectrum is broken into thousands of narrow bands and the relative absorbtivity at each band is recorded numerically on punched paper tape. The tape is then run through a special electronic computer that matches a prerecorded library of known spectrums to that of the unknown mixture to determine its chemical constituents and their concentrations. Only a few minutes is required for the complete analysis. The computer utilizes transistors and other advanced techniques such as plug-in printed circuits.

Direct-Voltage Instantaneous Breakdown of Oil-Impregnated Paper Capacitors as a Function of Area

By D. S. GIRLING

Standard Telephones and Cables Limited; London, England

THEORIES relating to the decrease of breakdown strength of dielectrics with increasing area are reviewed. The breakdown-voltage distribution of 2- and 3-layer capacitors of various areas and dielectric thicknesses are examined and it is shown that these are usually skewed towards lower values and may be double humped. The lower hump increases in importance with increasing area and is due to the coincidence of conducting particles in different layers.

It is shown that the variation of mean breakdown voltage with increasing area is represented adequately by Milnor's equation and that the breakdown voltage between spheres may be taken as the limiting value. Edge effect prevents such values from being approached with normal foil-type capacitors. The errors involved in the assumption of a normal distribution are assessed. The coefficient of variation appears to be independent of area but decreases with increasing thickness of dielectric. This information permits prediction of the voltage below which a given fraction of the sample will fail.

• • •

It has long been recognised that the breakdown voltage of thin sheets of dielectric decreases as the area increases. Small areas have breakdown voltages that are distributed around a relatively high value. Since failure of a larger area is due to the weakest of its constituent smaller areas, the breakdown voltages are distributed around lower values.

This electric strength variation is partly due to variations of the dielectric thickness and of the electric strength and partly to the inclusion of conducting or semiconducting particles generally arising from the tissue manufacturing, but sometimes also the capacitor winding process. Both these conditions result in electric stress variations from point to point in the capacitor area.

If the intrinsic electric strength of the material were uniform, failure in a uniform field would occur at the point of highest stress; that is to say, where the effective dielectric thickness is least. Thus, when failure is due to capacitor-tissue breakdown, the instantaneous breakdown voltage gives information about the minimum effective thickness. This may differ considerably from the nominal dielectric thickness. In connection with the rating of such dielectrics, a knowledge of this minimum effective thickness enables capacitors to be designed on a maximum-stress basis rather than on the nominal stress. It will be shown that on this basis, differing ratings are required for different dielectric areas or for different numbers of layers giving the same total thickness.

1. *Summary of Theory*

Although the reasons for the decrease in breakdown voltage with increasing area have been appreciated qualitatively for a long time, no completely satisfactory method of predicting breakdown voltage has been established. Milnor[1] and Holmes[4] produced equations based on the assumption that the breakdown distribution function of small areas was Gaussian. The work of Milnor was confirmed experimentally over a small range of areas by Bush and Moon[2].

Brooks[9] suggested that deviations from the true dielectric strength were due to conducting particles that reduced the effective dielectric thickness. After considering the mathematical probability of conducting-particle occurrence in relation to his experimental results, he showed that the conducting-particle sizes were distributed exponentially and that values greater than the paper thickness occurred. This resulted in the particle penetrating more than one layer of tissue. He concluded that it was not necessary to assume that coincidence between particles in the

[1] All footnote references will be found in the bibliography, section 5.

various layers occurs to cause such a condition and, further, it was possible for large conducting particles to become reoriented by such flexing of the sheet as occurs in winding.

In a further paper, Epstein and Brooks[10] applied the theory of extreme values to determine mathematically the effect of an exponential distribution of conducting-particle size on the average and on the distribution of breakdown voltage of capacitors of various areas.

Zingerman[12] gave a statistical theory and established an equation relating the breakdown-voltage distribution of large areas to that of a small area based on a generalised distribution of breakdowns of the small area. He pointed out that this distribution could not be Gaussian since his equations predicted that the breakdown distribution of larger areas would be asymmetrical. Since the small area itself could be regarded as a multiple of smaller areas, the small-area distribution must also be skewed.

If the probability that failure of a small area will occur at a voltage between x and $x + \delta x$ is $f(x)$, then the probability that a sample will break down at a voltage equal to or smaller than x is

$$F(x) = \int_{-\infty}^{x} f(x)\,dx. \qquad (1)$$

It therefore follows that the probability that n such areas will withstand a voltage x is

$$F_n'(x) = [1 - F(x)]^n. \qquad (2)$$

The breakdown voltage distribution is then

$$f_n(x) = dF_n'(x)/dx. \qquad (3)$$

and the mean breakdown voltage

$$\bar{x}_n = \int_0^{\infty} F_n(x)\,dx. \qquad (4)$$

1.1 Effect of Conducting Particles

In the experimental results described in the following paragraphs, observations were made over a large range of areas and it was clear from the histograms that the particle-coincidence effect must be considered. The conclusions of Brooks[9] are thought to be due to the assumption that all deviations from the nominal tissue strength were attributable to conducting particles.

The effect of conducting particles embedded in the tissue and penetrating it completely will be to produce a coincidence effect so that a capacitor can be short-circuited if particles in all layers are coincident or partially coincident. In the practical case, the sizes of the conducting particles will be distributed so that the breakdown-voltage distribution of a small area consisting of n tissue layers will consist of a series of distribution functions corresponding to $0, 1, 2 \cdots n$ coincident conducting particles. The distribution functions will have to take account of:—

A. The inherent variability of the electric strength of the dielectric material (including the effects of density and inhomogeneities other than conducting particles, et cetera).

B. The distribution of dielectric thickness.

C. The distribution of conducting-particle size in the direction normal to the paper.

D. The relation between thickness and breakdown voltage.

In voltage breakdown tests on very-small areas, the probability of conducting-particle occurrence will be so small that tests may not reveal any deviation from a single distribution. Combining small areas to form larger ones increases the probability of the occurrence of lower breakdown values, so that even in the absence of conducting particles, a decrease in electric strength would be observed. The lesser probabilities of failure, however, increase in proportion to the area ratio, so that the breakdown histogram becomes increasingly skewed towards lower values and may exhibit two maxima—the so called double-humped distribution. For normal tissue areas, the frequency of conducting-particle occurrence is such that there will be one somewhere in every layer so that the predominant distribution function will be that corresponding to 1 conducting particle. For larger areas, the occurrence of 2 coincident conducting particles in different layers may give rise to a lower hump. If the area were increased still further, or if the frequency of conducting-particle occurrence were

sufficiently high, the lower hump would become predominant and eventually a further lower hump due to 3 particles would become apparent.

2. *Experimental Results*

The work that is described in the following paragraphs was undertaken originally to ascertain empirically the variation of electric strength with area of a British-made rag-based tissue impregnated with mineral oil. The tests were carried out on groups of units varying from 50 to 500. The units were, in general, tubular for capacitances up to 0·1 microfarad and above that were of the flattened and clamped construction. The smaller tubular units and all the flattened units had tinned copper connecting tapes. The flashover gap at the edges normally varied from $\frac{1}{8}$ to $\frac{1}{4}$ inch (3·2 to 6·4 millimetres) according to the dielectric thickness. The paper width was normally in the range of 1 to 3 inches (2·5 to 7·6 centimetres) and was determined by the availability of supplies. This policy was adopted since it enabled the use of standard units and also because it was desired to obtain information on the units most commonly employed.

The units were impregnated in a mineral oil using a standard vacuum impregnation cycle; after impregnation they were cooled and stored under oil. Testing was carried out as soon as possible after impregnation to avoid any deterioration due to absorption of gas or moisture by the oil. Breakdown testing was carried out under oil. The following tests were made.

A. Quality-control check on capacitance, power factor, and insulance to ensure that impregnation was satisfactory.

B. Instantaneous-breakdown voltage measurement as described below.

C. Examination to eliminate failures due to faulty manufacture (torn papers, et cetera) to determine the nature of and position of failure and to check winding length. If, on examination, failure was proved to be due to faulty winding or mechanical damage to the paper, the observation was excluded.

In calculating the results of the tests described in these paragraphs, all valid observations were included.

In the course of this work it became apparent that the results obtained were consistent with one another and were forming a pattern that agreed qualitatively with the theory already given. The tests using sphere gaps were carried out to determine the limiting value for very-small areas.

2.1 Apparatus

The apparatus for measurement of breakdown voltage was a normal half-wave rectifier set; its input could be controlled by an adjustable-voltage transformer. The output was limited by a 1-megohm resistor. Indication of the breakdown voltage was obtained from an electrostatic voltmeter and the rate of increasing voltage was such that failure occurred in 5 to 10 seconds. Since observations were made on a rising voltage, each observation was liable to some error. It has not been possible to evaluate this, but it is clear that such an error will be nonsystematic and will limit the minimum standard deviation that can be observed.

2.2 Tests for Conducting Particles

Analysis of the results of tests for conducting particles, using the roller method described in British Standard Specification *698* of 1936, shows that the number of particles that completely penetrate the paper decreases rapidly with increasing thickness. This test cannot be regarded as absolute since it depends on the flatness of the plate and the uniformity of the roller; great care must be taken to eliminate dust. An assessment of the effectiveness of the apparatus can be made by carefully testing a piece of paper by making several runs with it and marking all detectable conducting particles. It has been found that in one run at least 60 per cent of all detectable conducting particles is indicated. The number of such conducting particles as a function of thickness is shown in Figure 1.

2.3 Sphere-Gap Tests

To measure the instantaneous-breakdown stress of the smallest practicable area, a series of tests was carried out using a test cell made to British Standard Specification *148* of 1951.

The test results are shown in Figure 2 on which the breakdown stress in volts per micron for 50 tests are plotted as a function of total thickness.

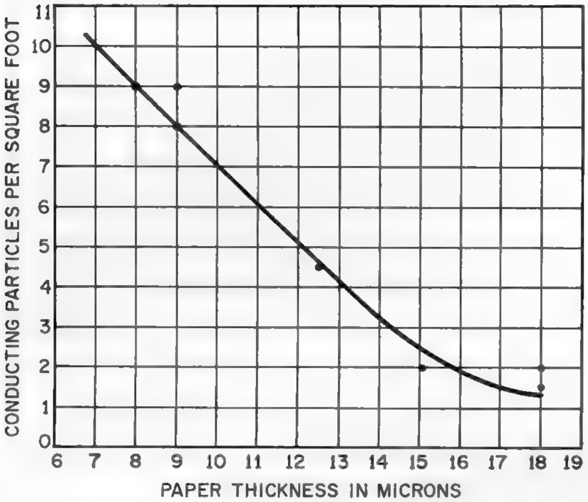


Figure 1—Number of conducting particle per square foot as a function of thickness for rag-based tissue. (1 square foot = 0·093 square metre.) Each point represents the average of 24 1-by-12-inch (2·54-by-30·48-centimetre) samples. Measured in accordance with British Standard Specification *698*.

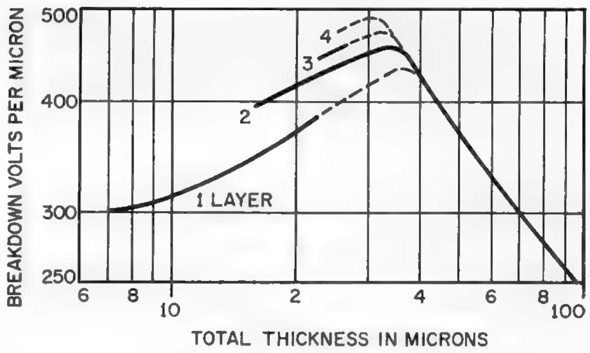


Figure 2—Instantaneous direct-current breakdown voltage of rag-based tissue impregnated with mineral oil. Test was between 0·5-inch (1·25-centimetre) spheres at 20 degrees centigrade.

Depending on the number of layers, increasing thickness of tissue results in increasing strength up to a certain critical value. Increasing the number of layers also results in increased strength. Above the critical value this stress decreases again and, as far as can be seen, is independent of the number of layers. The dependence of breakdown stress on the number of layers has also been observed by Hopkins, Walters, and Scoville[14]. Since the breakdown stress is a function of the number of papers below the critical value, and not above, it is considered that below this value failure is inititated by the paper and above it by the oil. Examination of the failure distribution shows that as far as can be judged on groups of 50 results, they follow a Gaussian law.

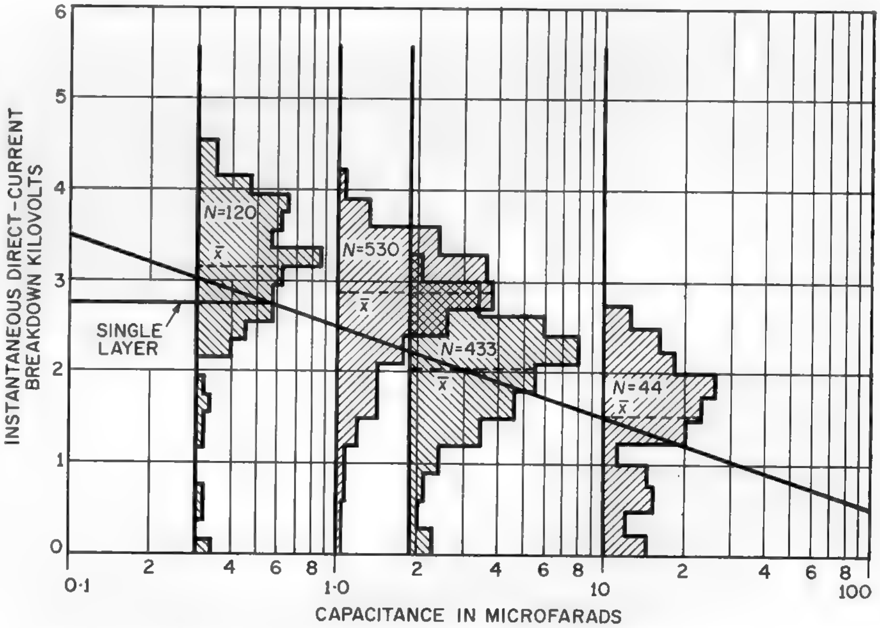


Figure 3—Instantaneous direct-current-breakdown-voltage histograms as a function of capacitance (area). Four values of capacitance, 0·3, 1, 2, and 10 microfarads, were examined. All had 2 layers of 9-micron rag-based tissue impregnated with mineral oil. N = number of capacitors represented in each histogram.

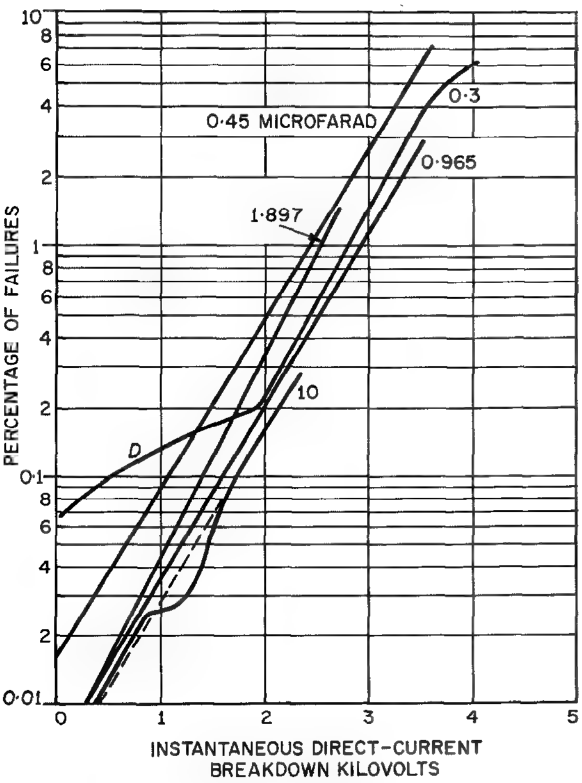


Figure 4—Breakdown voltage of capacitors consisting of 2 layer of 9-micron rag-based tissue impregnated with mineral oil, referred to 0·01 microfarad. The portion of the 0·3-microfarad-capacitor curve marked *D* is a deviation caused by other types of failure; these were the only completely assembled capacitors in these tests.

2.4 Tests on Wound Units

2.4.1 Breakdown Histograms

2.4.1.1 Double-Layer Capacitors

The test results on units wound with 2 layers of 9-micron tissues are shown in Figure 3. This is drawn as a plot of instantaneous-breakdown voltage against capacitance; superimposed on the plots are histograms showing the failure distribution. Two of these are based on 433 and 530 units and it can be seen that these present a smooth and slightly skewed appearance. These 4 distributions demonstrate fairly clearly the points suggested by the theory.

The 1-microfarad-capacitor results show a slight skewness towards lower breakdown voltage values. The increase to 2 microfarads shows a similiar distribution about a rather lower value and a marked increase in the failures in the lowest cell. This trend continues still further with the increase to 10 microfarads. The strength of a single-layer 9-micron tissue can be obtained from Figure 2 as 2·75 kilovolts. Since in the 0·2-microfarad capacitor there will be on the average about 4 conducting particles in each layer of paper, it may be assumed that in most capacitors, only one layer will be effective. The probability of two coincident conducting particles is still too small to be noticeable.

Figure 4 shows the failure distribution for 0·01-microfarad capacitors calculated from the above results using (2). If the breakdown voltage were a function of area only, then all these curves would coincide. It is quite clear that they tend to do so but the extreme cases show a discrepancy by a factor of 3 in the failure percentage. It is interesting to note that what appeared as a double hump in the 10-microfarad-capacitor experiment does not appear to be so by comparison with the other curves. The divergence of the results for the 0·3-microfarad capacitor (the only complete capacitor tested) at the lower voltages clearly demonstrates the existence of faults resulting from causes other than dielectric failures. Figure 5 shows the percentage of failures

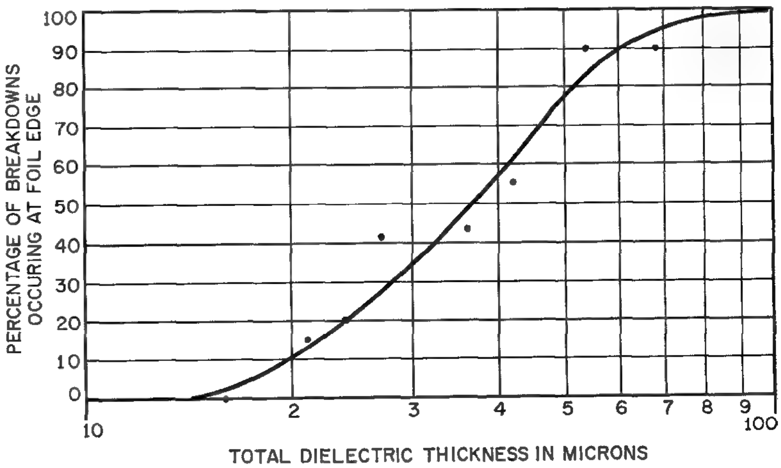


Figure 5—Relation between percentage of failure occurring at foil edge and total dielectric thickness. Plotted points are average values.

that occurred at the paper edge in tests on capacitors of various dielectric thicknesses and demonstrates the increasing importance of edge effect with increasing thickness.

2.4.1.2 Triple-Layer Capacitors

A set of results similar to Figure 3 is shown in Figure 6 for the case of 3 layers of 8-micron tissue. These are shown over a wider area range. The

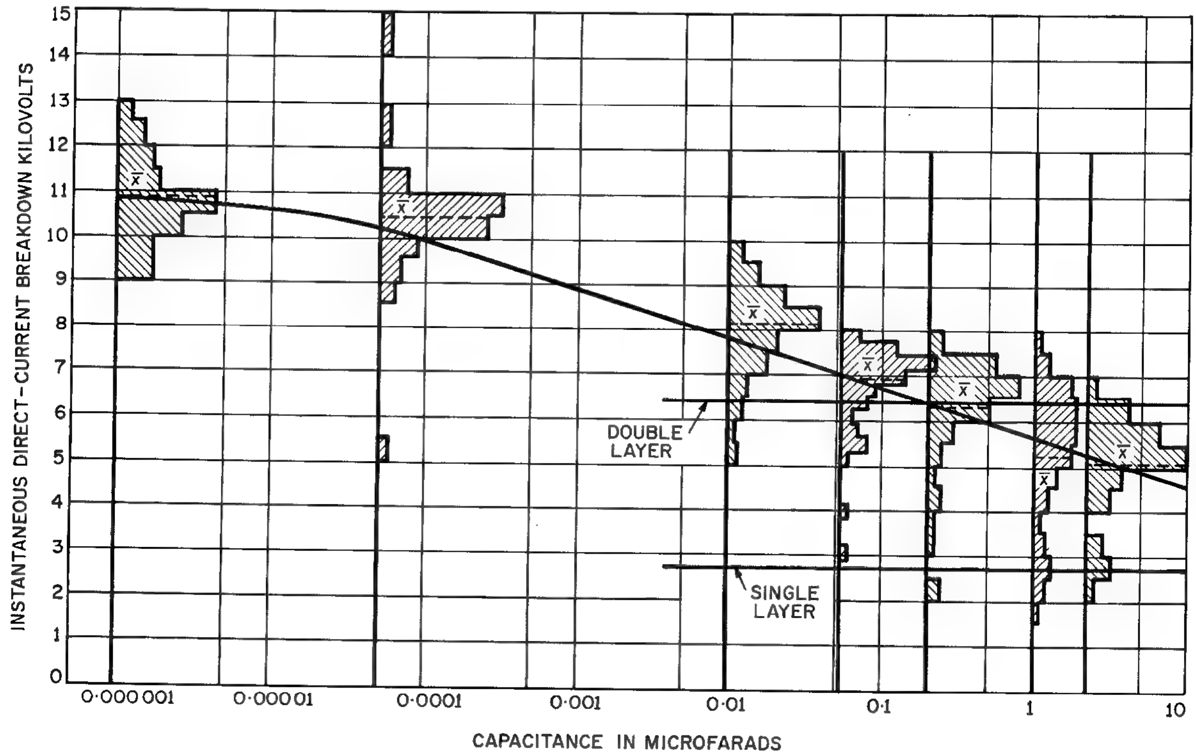


Figure 6—Instantaneous-breakdown-voltage histograms as a function of capacitance (area). Values of capacitance tested were 0·000 001 (sphere gap), 0·0001, 0·01, 0·05, 0·2, 1, and 3 microfarads. All had 3 layers of 8-micron rag-based tissue impregnated with mineral oil. Each histogram is based on 100 results.

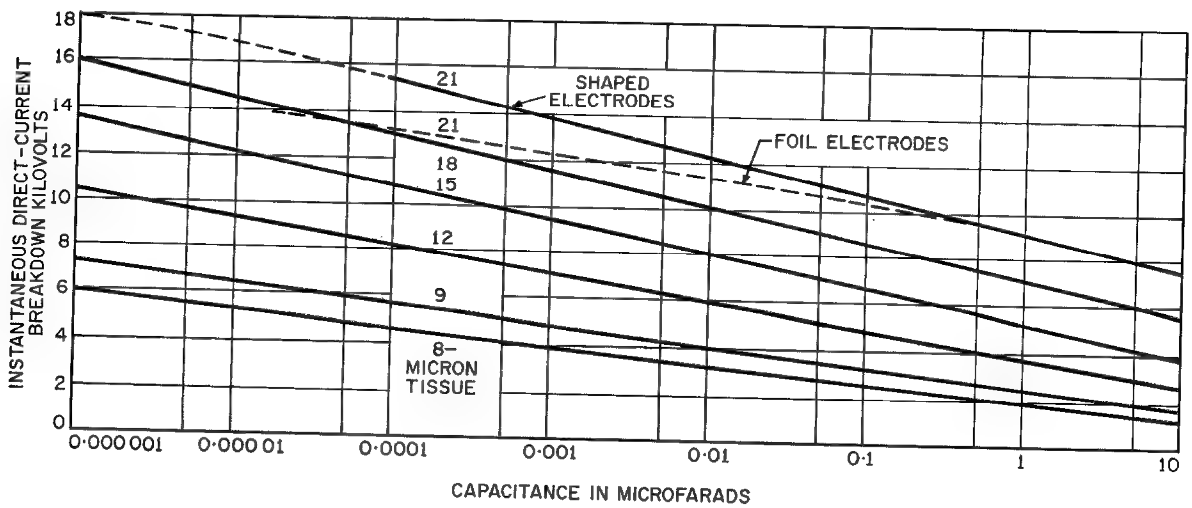


Figure 8—Breakdown voltage versus capacitance for 2 layers of rag-based tissue impregnated with mineral oil at 20 degrees centigrade.

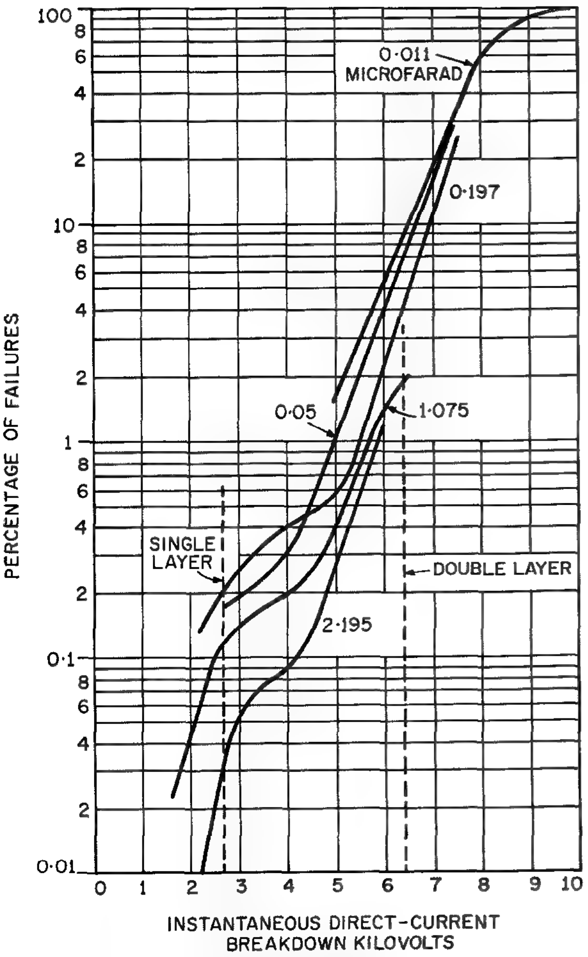


Figure 7—Breakdown voltage of capacitors consisting of 3 layers of 8-micron rag-based tissue impregnated with mineral oil, referred to 0·01 microfarad.

point plotted at 0·000 001 microfarad is for the sphere-gap test. Once again the trend is downward as the area increases. For the sphere-gap tests, no failures occurred outside the main distribution, but as the area increased there was an increasing number of such failures. The failure distribution for the 1-microfarad capacitor is one of the few clearly double-humped distributions that have been observed. It will be noted that the lower hump is centered on the strength of a single layer of tissue as determined by the sphere-gap tests.

Figure 7 shows the cumulative failure distribution for these results when referred to a 0·01-microfarad capacitance by use of (2). As for the 2-layer case, the curves should coincide if breakdown voltage is a function only of area. A second hump is apparent in all tests except that in which failures were concentrated in the upper hump. The lower hump clearly corresponds to the failure of a single layer of tissue. At 0·01 microfarad, about 0·1 per cent of the sample has one effective paper.

2.4.2 *Correlation of Results*

2.4.2.1 *Average Values*

The results of a large number of tests on different areas and different tissue thicknesses are shown in Figures 8 and 9 for the 2- and 3-layer cases, respectively. The points plotted on these graphs are the average values of at least 50 results. The test results using sphere gaps have

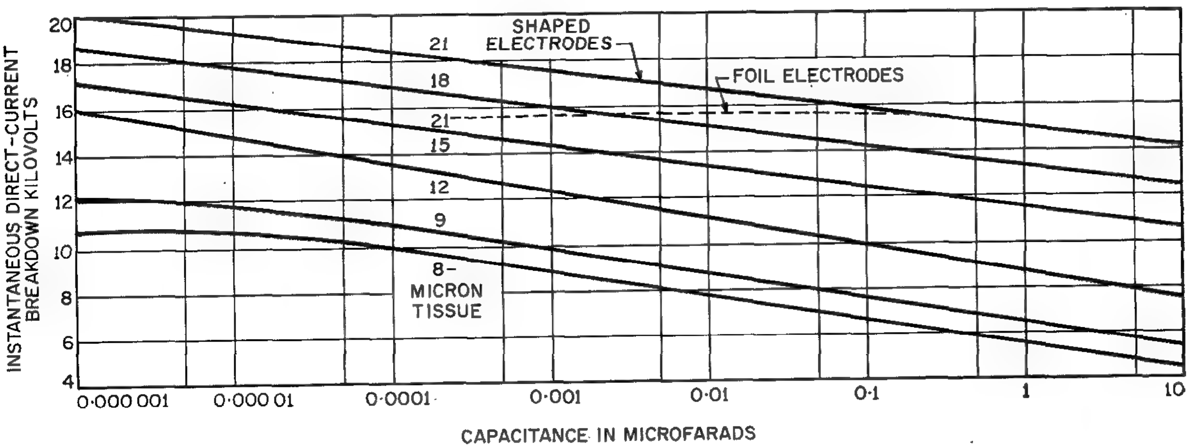


Figure 9—Breakdown voltage versus capacitance for 3 layers of rag-based tissue impregnated with mineral oil at 20 degrees centigrade.

been assumed to be the small-area limiting values and these are plotted at a 0·000 001-microfarad capacitance. A family of lines has been constructed using horizontal and vertical interpolation and, although the two sets of lines were independent of area. Tests on 21-micron tissue show that with foil electrodes, there is a marked limiting value due to stress at the edge. Tests with shaped electrodes gave only slightly higher values and did not reach the line interpolating

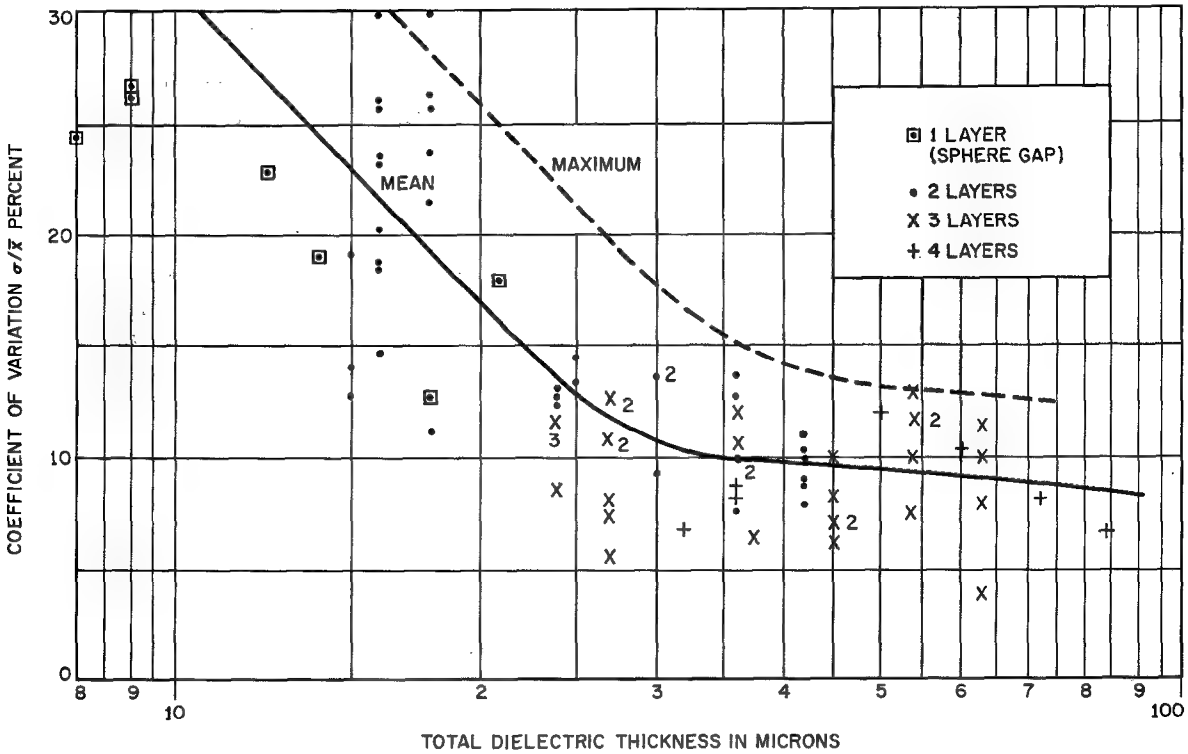


Figure 10—Coefficient of breakdown-voltage variation versus dielectric thickness of various areas of rag-based tissue impregnated with mineral oil. Points plotted are each the results of tests on 50 to 100 units.

obtained separately, it can be seen that their slopes are very similar and for the most part are parallel. For the 2-layer case with thicknesses below 12 microns, there appears a slight tendency for the lines to converge as area increases.

In the case of 2 layers of 21-micron tissue using foil electrodes, the curve tends to flatten for decreasing area and does not approach the sphere-gap limiting value. In such tests 90 per cent of the failures occurred at the edge and it was apparent that the stress at this point was the limiting factor. Tests using shaped electrodes reduced this value to about 60 per cent and resulted in a higher average value.

In the 3-layer case, it is not possible to draw a straight line covering the whole capacitance range for the thinner thicknesses and therefore the small-area breakdown strength appears to be between sphere-gap measurements and those made on large areas.

The graphs are drawn on a log-linear scale and hence the law obeyed is

$$E = A - B \log C, \tag{4}$$

where A and B are constants. This is in the form of Milnor's equation.

2.4.2.2 *Coefficient of Variation*

Examination of the coefficient of variation calculated in the experiments described previously shows that, although there was some variation from one batch to the next, this was not related to increasing area. The most-reasonable assumption is that this value is independent of area. A scatter diagram of the results is shown

ı Figure 10, in which the variation coefficient is lotted against total dielectric thickness. As the nickness increases, it tends towards a value of bout 10 per cent. This value will include some xperimental error.

.4.2.3 *Skewness*

It is apparent that while the theoretical failure-requency distribution consists of a number of uperimposed distributions, each of which may ır may not be skewed, tests on small numbers of amples will reveal only a skewed distribution ather than a double hump. The result is that nore failures occur at low voltages than would ›e expected from a knowledge of $\bar{x}$ and σ. The ›ercentages of failures that occurred at $(\bar{x} - 1\cdot65\,\sigma)$, $(\bar{x} - 1\cdot3\,\sigma)$ and $\bar{x}$ corresponding .o 5, 10, and 50 per cent respectively are plotted n Figure 11, which shows that for the thinnest lielectrics the actual failures at a voltage that ;hould give 5-per-cent failures may, in fact, be as nuch as 12 per cent. The error is less for predic-:ion of larger failure percentages and for the :hicker dielectrics and becomes less as the dis-:ribution skewness decreases. This information can be used for rejection-rate predictions on voltage proof test.

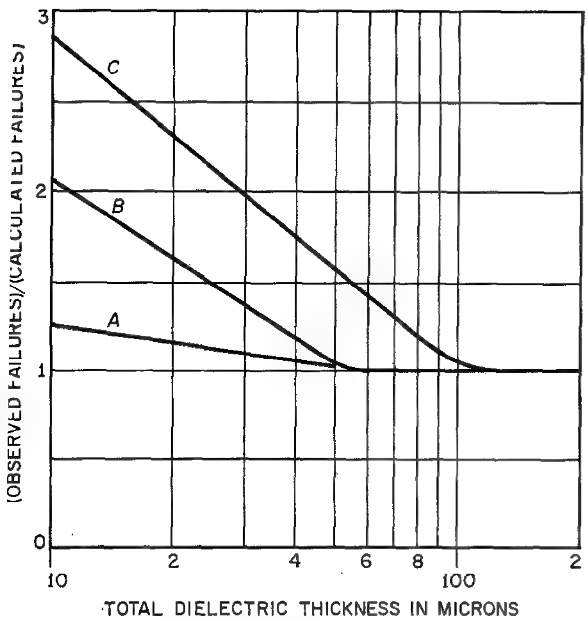


Figure 11—Calculated and measured failure rates versus dielectric thickness, plotted from averages of measurements on various areas. Curve A is the ratio for $\bar{x}$ (50-per-cent failure), B is for $\bar{x} - 1\cdot3\sigma$ (10-per-cent failure), and C is for $\bar{x} - 1\cdot650\sigma$ (5-per-cent failure).

3. *Conclusions*

A. In tests on capacitor tissues placed between 0·5-inch (12·7-millimetre) spheres, the breakdown stress increases with tissue thickness up to a certain critical value, above which it decreases. Below this value, increased strength is obtained for any one total thickness by increasing the number of tissues. Above it, the strength appears independent of the number of tissues and it is possible that this is due to the oil.

B. Mean-breakdown-voltage variation of 2-layer capacitors with area is represented adequately by Milnor's equation and sphere test results appear to be the limiting values on the assumption that they correspond to a capacitance of 0·000 001 microfarad. Deviation from this law results from edge-effect, which becomes increasingly important as thickness increases. This prevents the limiting values observed from sphere tests from being achieved with foil-type capacitors.

C. The mean breakdown voltage of 3-layer capacitors follows a similar law although the thinner tissues depart from the law, probably due to the alignment effect between conducting particles embedded in the tissues, causing breakdown distributions to become double-humped. Owing to the increased dielectric thickness, the edge effect is much-more important. As pointed out by Zingerman[12], this is a function of the edge length and not of area. The geometry of the unit must therefore be considered.

D. Breakdown-voltage comparison of 2- and 3-layer 1-microfarad capacitors of 24-micron total dielectric thickness shows that the 3-layer capacitors have a strength 50 percent greater than the 2-layer.

E. Although the coefficient of variation is itself quite variable, it appears to be independent of area and decreases with increasing thickness. The failure distribution is not normal, except possibly for the sphere tests, and the distributions are skewed towards lower values. The assumption of a normal distribution is, however, a useful one

and the errors involved have been established empirically. By this method, the fraction of a sample that will fail below a certain voltage can be predicted.

F. Prediction of the large-area breakdown-voltage distribution from tests on small areas over large ranges is liable to error unless a very-large number of samples is used; in which case it is simpler and just as economical to test the larger areas. The construction of the graphs shown permits a ready estimate of the mean breakdown voltage and coefficient of variation of any intermediate area. Owing to the coincidence effect of conducting particles, extrapolation should not be carried out over large ranges.

4. *Acknowledgments*

The author records his appreciation of the advice and encouragement given by Mr. R. M. Barnard, without which this paper would not have been possible, and the assistance of Messrs. W. E. R. Evans, A. Simmons, and R. S. W. Walker.

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Characteristics and Applications of the Iatron Storage Tube*

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THE OPERATION of many types of storage tubes depends on intensity modulation of an electron beam by an electrostatic charge on an insulator layer.[3–7] As early as 1927, Dr. P. T. Farnsworth suggested the method as a means of increasing brightness in cathode-ray tubes for television.

Iatron® storage tubes have been under development since 1949 and have been in experimental use since 1953. The properties of image storage and extremely high brightness at low voltage make the tube very attractive for radar indicators, oscilloscopes, and other uses.

This paper describes Iatron operation with emphasis on those unusual characteristics that have no counterpart in ordinary cathode-ray tubes and suggests modifications of an oscilloscope to use the tube. It is hoped that the applications engineer will find the answers to some of the questions that may arise when operating this storage tube for the first time.

1. Description

The Iatron is a storage cathode-ray tube in which the display can be written, stored, and viewed continuously. While it is available in a variety of bulb sizes and types of writing guns, the characteristics presented here are particularly applicable to the 5-inch (127-millimeter) *Ia10P20-25* tube, shown in Figure 1, which has an electrostatic writing gun. Continuous viewing

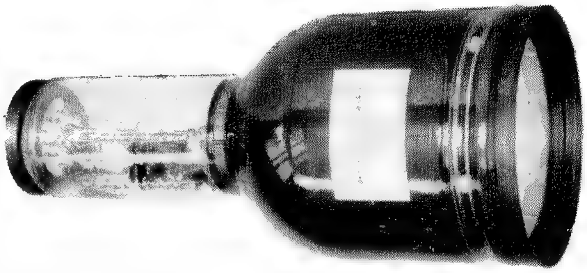

Figure 1—The 5-inch (127-millimeter) electrostatic-deflection Iatron *Ia10P20-25*.

is provided by a divergent electron beam, called the flooding beam, which expands until it covers the entire display area of the tube. Elemental areas of the large flooding beam are modulated by a conventional cathode-ray beam.

The following explanation of the method of modulating the flooding beam refers to Figure 2. The flooding gun is located on the tube axis at the juncture of the neck and bulb. The control grid of the flooding beam is an insulating layer that covers the gun-facing side of a fine-mesh metallic screen. This insulator-screen is located in the path of the beam at the distance where the beam has expanded to its maximum size. The beam passes through the metallic screen and impinges on the aluminized phosphor on the inner face of the tube. An aquadag coating on the wall of the bulb and a metallic collector screen serve to collimate the flooding beam. With proper collimation, the paths of the electrons in the expanded flooding beam are made parallel to each other and perpendicular to the plane of the insulator-screen.

The metallic screen that supports the insulator layer serves the same purpose as the screen grid of a tetrode and is operated on a fixed voltage

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* Reprinted from *Communication and Electronics*, number 29, pages 47–53; March, 1957. Iatron is a registered trademark of Farnsworth Electronics Company. Dr. P. T. Farnsworth has made many notable contributions to the storage-tube art[1,2] and has personally directed the Iatron research and development work undertaken in 1949. Others who have contributed to the development of the Iatron include a majority of the Farnsworth research department. Current Iatron development has been supported primarily by the United States Navy Bureau of Ships.

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[6] United States Patent 2 228 338.

[7] United States Patent 2 754 449.

of about +10 volts. Points on the insulating surface, however, can assume potentials that are quite different from each other and from the voltage on the metallic screen. These potentials fade to cutoff. Without them, the entire display area would eventually increase to maximum brightness. The increase in brightness results when positive ions, which are generated by colli-

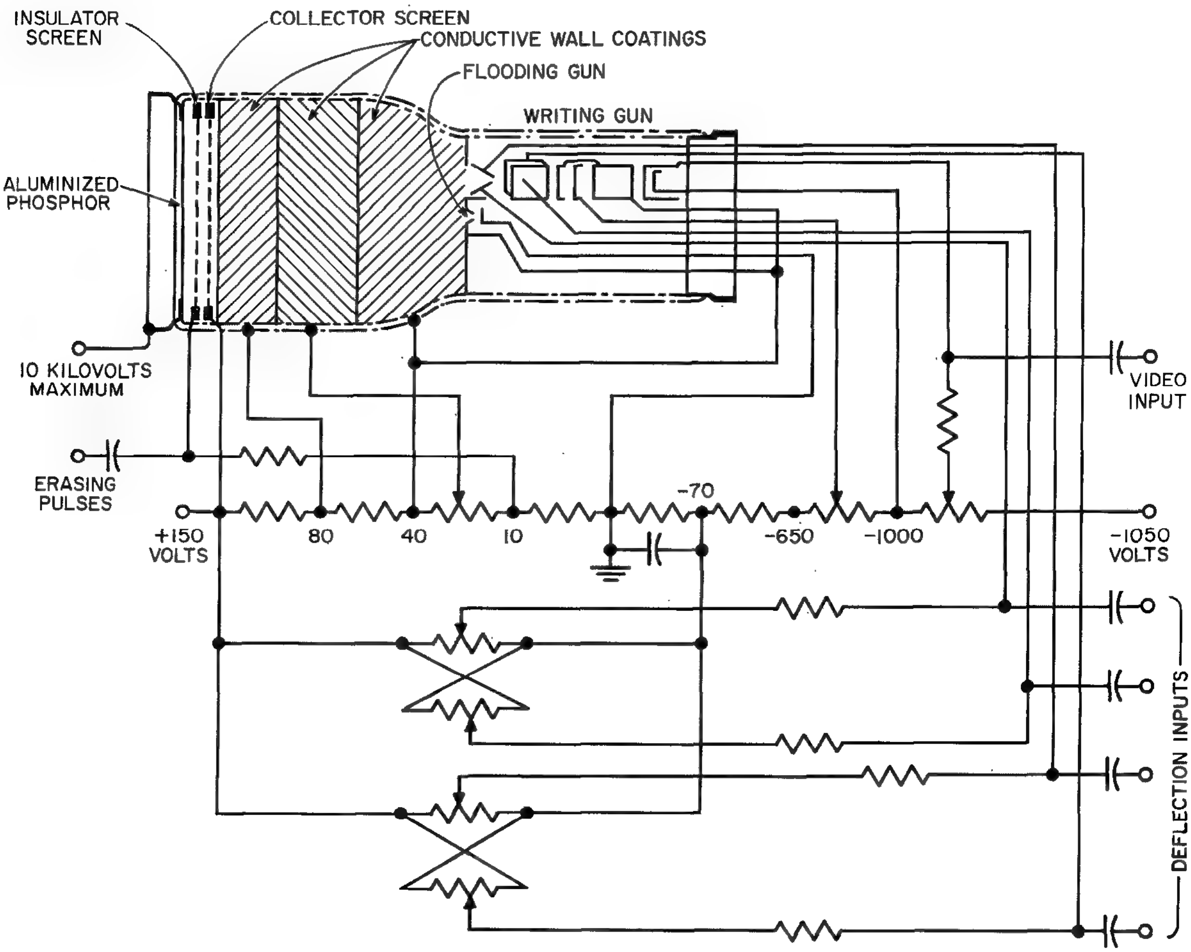


Figure 2—Diagram of Iatron model *Ia10P20-25* and associated circuits.

are established on the insulating surface by charging it with the writing beam and the flooding beam. Over the insulator control range, the number of flooding electrons that can pass through the screen at any point is roughly proportional to the potential at that point of the insulating surface.

1.1 Erasing

Erasing pulses are essential to the operation of the Iatron. These are low-voltage pulses applied to the insulator-screen that cause the display to

sion between flooding electrons and residual gas molecules, are drawn to the insulator surface, charging it positively. The ion density over the insulator surface is uniform and therefore the display area brightens everywhere at the same rate.

Further increase of insulator potential ceases when the insulator has become charged to about +2 volts with respect to the flooding-gun cathode. At this potential, some flooding electrons have enough energy to strike the insulator. Striking with low energy, these electrons adhere and, by virtue of their negative charge, prevent addi-

tional voltage increase. An equilibrium insulator potential is thereby established at which the positive ions are balanced by the electrons landing on the insulator.

When the insulator suddenly becomes more positive, as by applying a voltage pulse to the support screen, flooding current strikes the insulator, quickly charging the surface down to equilibrium again. At the end of the pulse, the support screen is returned to its original potential, but the insulator is now more negative than it was before because of the charge acquired during the pulse. Positive insulator charges written by the writing beam are erased in the same way.

1.2 Insulator Control Characteristic

Figure 3 is a curve showing average phosphor flooding current over the display area as a function of insulator surface potential. The points on the curve were obtained, using a +10-volt bias on the support screen, by first allowing ions to charge the entire insulator surface to equilibrium. At equilibrium, the insulator is charged to its most positive potential and phosphor current is a maximum. This current corresponds to zero insulator volts on the curve. The insulator surface was then charged 0.5-volt negatively by applying a +0.5-volt pulse to the insulator support screen. The phosphor current was measured at the end of the pulse and the process was repeated, increasing the amplitude of the pulse at each step until phosphor current reached cutoff. Thus it was determined that pulses of 3.2-volt amplitude are necessary to erase the tube to cutoff. The assumptions were made that the insulator surface is at zero volts at equilibrium and that the insulator is charged negatively to the amplitude of the erase pulse at each step. Although, as a matter of academic interest, the insulator surface potential at equilibrium was actually about 2 volts, the first assumption is valid in the sense that the insulator does not draw flooding current over the control range from equilibrium to cutoff.

1.3 Control of Viewing Time

It is clear that the insulator cannot be charged to a potential more negative than the absolute amplitude of the erasing pulse, because it charges to an equilibrium potential that is reached when the voltage difference between the flooding-gun cathode and insulator is just sufficient for flooding electrons to strike the insulator. But it can be prevented from charging to this equilibrium if the pulse duration is shorter than the charging

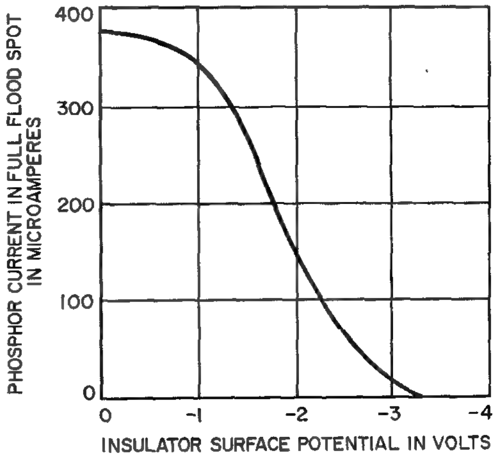


Figure 3—Static insulator control characteristic.

time required. The erasing pulses used to obtain the curve of Figure 3 were applied for a sufficient length of time to charge the insulator to equilibrium.

The average time required to charge the insulator from maximum brightness to cutoff is about 3 milliseconds and a display can be made to persist for many seconds if pulses that are narrower than 3 milliseconds are applied continuously at a suitable repetition rate. On each successive pulse, the insulator will be charged negatively a small amount and will eventually reach cutoff when the product of pulse width times the number of consecutive pulses is equal to 3 milliseconds.

Pulses having a repetition frequency of f pulses per second, which are t seconds wide and 3.2 volts in amplitude, will therefore erase signals from maximum brightness to cutoff in the approximate time $T = 0.003/ft$ second, for large values of f. (The erasing-pulse amplitude required to erase to cutoff is increased slightly if the erasing-pulse frequency is reduced to very-low values, since the insulator voltage shift by ions during the interval between pulses then becomes significant.) Hence, the viewing time of the display can be varied by controlling either the frequency or the width of the erasing pulses. The proper choice

of pulse repetition frequency is influenced by flicker and contrast as well as by viewing time.

1.3.1 Flicker

During an erasing pulse, the entire display area is illuminated by flooding current which passes through the insulator screen (see Figures 5 through 7). The resulting flashes of light produce flicker at low pulse frequencies, or merge to a constant background brightness at a pulse frequency of about 45 pulses per second. However, very-narrow pulses such as would be used to obtain a viewing time of the order of 1.0 minute are hardly detectable in the display even in the flicker-frequency range.

1.3.2 Contrast

Loss of contrast is caused by background brightness. The average background contributed by erasing pulses is $B = 0.003/100T = 100\,ft$ percent of maximum display brightness.

Loss of contrast would be most severe using the erasing conditions required for minimum viewing time. However, this is an unrealistic condition for the Iatron as it defeats the very purpose of storage tubes. Nevertheless, it is interesting to estimate the maximum background brightness that might be encountered.

Since the minimum erasing time is 3 milliseconds and the repetition rate over which persistence is limited by the eyes of the observer is 45 pulses per second, these values may be taken as the pulse width and frequency that will determine minimum viewing time. Under these conditions, the display area would be viewed at maximum brightness $45 \times 0.003 = 0.135$ second per second of total viewing time. The average background brightness would, therefore, be about 13.5 percent of the maximum signal brightness.

However, by the same reasoning, B would be only 0.3 percent for a viewing time of 1.0 second. At 45 pulses per second, the pulse width for this condition would be about 67 microseconds.

1.4 Maximum Viewing Time

Viewing time can be limited by positive ions or by insulator leakage. Leakage is negligible at normal tube operating temperatures and most Iatrons will store written charges for several hours if all tube voltages are removed after writing, to avoid generation of ions.

The number of ions generated is a function of several things such as flooding-current density, length of the electron paths, pressure of residual gases in the tube, et cetera. The time required in an average tube for ions to charge the insulator from cutoff to 50 percent of maximum brightness is about 30 seconds. The measurement is significant in that there must be a minimum product of erasing frequency and pulse width, ft, which will prevent ion charge integration on the insulator and thereby maintain nearly constant insulator potential in the interval between pulses.

Figure 4 shows curves of phosphor current versus time for a typical tube using small values

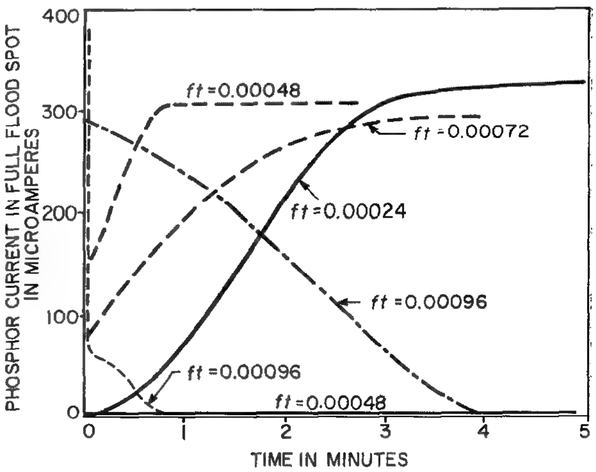


Figure 4—Change of phosphor current caused by positive ions and erasing pulses. Initial conditions: Solid line—phosphor current initially cut off. Dashed line—phosphor current initially maximum. Dash-dot line—phosphor current initially 330 microamperes. Parameters: Erase pulse width t seconds. Erase pulse frequency $f = 60$ pulses per second.

of ft. The ordinate is proportional to brightness since the flooding-spot size is constant and the power input to the phosphor is well below the level at which saturation occurs. For the condition that the insulator is initially at cutoff, an erasing product $ft = 0.000\,48$ maintains cutoff. If the insulator is initially at equilibrium (phosphor current initially maximum), however, an erasing product $ft = 0.000\,48$ will not erase the tube but will allow the insulator to charge to an intermediate potential corresponding to 310 microamperes of phosphor current. Thus it is found

that two stable potentials, one corresponding to cutoff and the other corresponding to 310 microamperes of phosphor current, are possible. They are maintained in each case because the insulator charge received from the flooding electrons during one erasing pulse is equal to the ion charge acquired during the interval between pulses.

A necessary condition for continuous operation is that signals written by the writing beam to any brightness will eventually be erased to cutoff. Lacking sufficient erasure, the display area would soon be saturated. As shown in Figure 4, if a minimum erasing product $ft = 0.000\,96$ is used, a signal that has been written to maximum brightness will assuredly be erased.

Two curves are shown for which $ft = 0.000\,96$. In the first, the pulse was not applied until the insulator had been charged to equilibrium and allowed to remain there for a considerable length of time. By suddenly applying the pulses, phosphor current drops sharply and the tube erases to cutoff. In the second curve, erasing pulses were not applied suddenly but were present continually while the tube was written to the initial phosphor current of 330 microamperes. This curve is typical of viewing-time curves obtained with continuously operating erasing pulses, the viewing time being less for larger values of ft.

The cause of the rapid decrease in brightness that is observed when an erasing pulse is suddenly applied, under the initial condition that the phosphor current is a maximum, is not fully understood. However, it is probable that the distribution of charges on the insulator on the front surface as well as on the sides of the mesh holes of the insulator screen is affected by initial conditions. Depending on the distribution of charges, the electric fields in the vicinity of the meshes will be modified, deflecting the flooding electrons and ions to selective minute areas of the insulator surface and, consequently, altering the erasing characteristic.

It is interesting to note that if the insulator is initially at cutoff and $0.000\,48 < ft < 0.000\,96$, insulator areas where a large writing charge has been deposited will charge to a definite brightness level, while weak signals will be erased to cutoff. (A small area of positive charge will not persist indefinitely however, since the escape of secondary electrons from the area is inhibited by the electric field of the less-positive surrounding insulator surface, thus causing more electrons to stick, charging the area negatively. The bright area shrinks in size and disappears.) This characteristic can be used to achieve extended viewing time but with concurrent loss of half tones. Extension of viewing time by this method is even greater if still-narrower erasing pulses of relatively high amplitude are used. The viewing time obtainable in this way is upward of 30 seconds, after which time the written areas decay rather rapidly to cutoff.

1.5 Writing Characteristic

With the tube operating and with proper erasing pulses being applied, the tube will be at cutoff and in readiness to be written on. The 1000-volt writing beam can be scanned over the insulator in any desired pattern and video signals are applied to the control grid to modulate the beam.

The collector screen, insulator, and phosphor each intercept a part of the writing beam. The current intercepted by the insulator serves to charge it in the positive direction since the beam energy is great enough to eject a greater number of secondary electrons than the number of primary electrons intercepted.

The rate at which the insulator can be charged by the writing beam is very high. This rate is determined by measuring the brightness of the stored trace after the writing beam has been deflected across the tube at a known scanning speed. In Figure 5, the brightness of a stored signal is plotted as it increases in stair-step fashion, with each passage of the writing beam during successive superimposed scans across the corresponding point on the insulator. The writing spot was scanning at a speed of 2.75×10^4 centimeters per second and the three curves are for three values of writing-beam current. After the tenth scan, an erasing pulse was applied to restore the insulator to cutoff. The bright flash which accompanies the erase pulse is shown also in Figure 5. It can be seen that the flash is never brighter than the brightness associated with equilibrium insulator potential and has a relationship to the brightness of the signal that is being erased.

The writing currents used to obtain the data in Figure 5 were necessarily very low because of the relatively low scanning speed available

for these tests. However, the writing-beam current can be as high as 150 microamperes at zero grid bias and will write to maximum brightness in a single-line scan at a speed of about 10^5 centimeters per second.

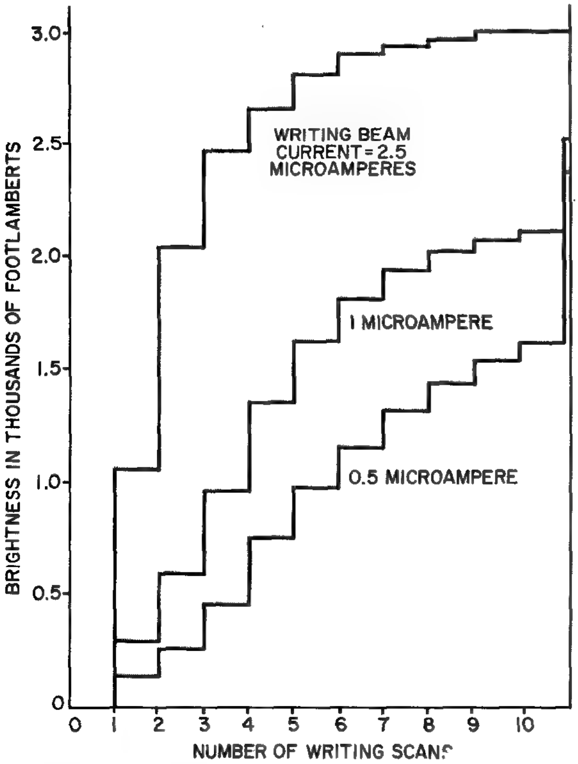


Figure 5—Brightness of a point written on for 10 successive writing scans and finally erased to cutoff; writing-spot scanning speed 2.75×10^4 centimeters per second.

At high brightness, the brightness increase per scan grows smaller. At the high-brightness limit, saturation is reached and further writing results only in a charge spreading on the insulator surface. Since an electron beam does not possess finite size, a few electrons will be found even at radii considerably larger than the dimensions of the spot defined by the usual methods. These fringe electrons continue to integrate on the insulator area surrounding the core of the spot after it has saturated.

The obvious implication is a serious one, that areas repeatedly written upon in a display will tend to bloom unless some form of insulator writing-charge limiting is used. In a plan-position-indicator radar display, for example, an equalizing signal derived from the range-deflection voltage can be added to the video signal to prevent blooming at the center. To emphasize weak targets in a display, video clipping will prevent blooming on strong signals. Proper choice of the erasing-pulse frequency is, of course, extremely important in this matter since writing charge can be erased before it has had the opportunity to integrate beyond a desired level. An erasing frequency equal to the deflection frequency is the best choice in many cases.

In almost any continuous display, the erasing-pulse frequency and pulse width are preset; after adjustments have been made to achieve the desired viewing time no further adjustments will be required during operation. Figure 6 illustrates typical continuous operation of an Iatron using an erasing frequency equal to the writing-beam scanning frequency, which in this case is 60 cycles per second. An erasing-pulse width of 0.001 35 second was used and the viewing time was consequently about 37 milliseconds. The way in which the average brightness varies with writing current is apparent from the three conditions shown. The two brightness levels in each graph

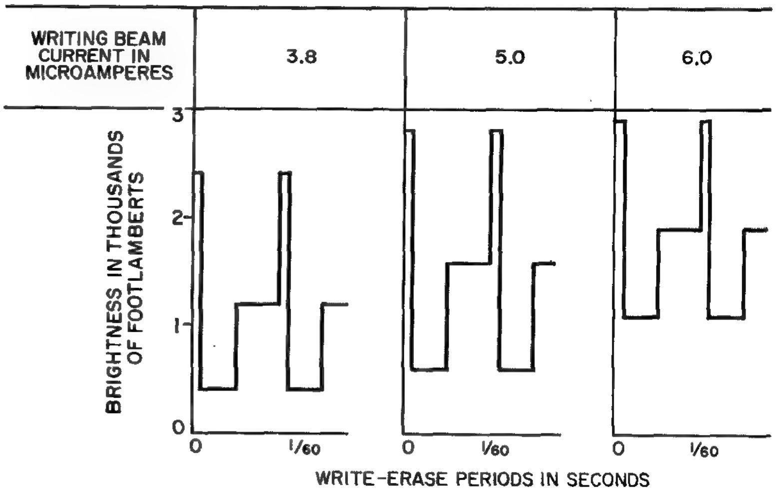


Figure 6—Stable brightness levels of a point alternately written and erased; erasing-pulse width = 1350 microseconds.

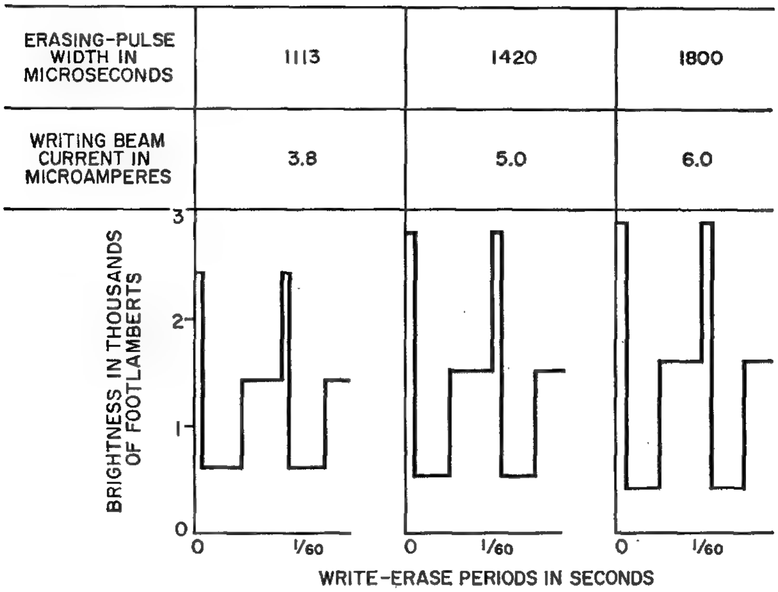


Figure 7—Stable brightness levels of a point alternately written and erased; average brightness is constant.

correspond to insulator potentials immediately after writing and immediately after erasing. These are stable potentials that will be repeated until either writing current or erasing-pulse width is changed. Each erasing pulse charges the insulator downward in potential by an amount equal to the potential increase by writing during one scan. The ability of the insulator to assume stable potentials follows from its nonlinear charging characteristic wherein the charging rate decreases toward higher brightness levels; Figure 5. If the characteristic were linear, writing charges would integrate to saturation. Since the erasing-pulse amplitude was 3.2 volts, the tube would have been erased just to cutoff if writing current were reduced to zero.

As indicated in Figure 7, a change of viewing time requires adjustment of both average writing current and the amount of erasure. In Figure 7, the average brightness was held constant by adjusting the writing current, after the viewing time was changed, by altering the erasing-pulse width.

1.6 Resolution

To measure resolution, a raster of a known number of equally spaced lines is scanned, and the raster is shrunk until the individual lines are no longer discernible. The raster width that is normal to the lines is then measured and resolution is specified in lines per inch.

Resolution measurements are meaningless unless brightness is specified concurrently. It is found that resolution is approximately the same at a given brightness regardless of how many scans or what writing-beam current were required to write to that brightness. While the resolution of a stored image at low brightness is nearly equal to the resolution of the writing beam itself, at high brightness it approaches a minimum of about 35 lines per inch (14 lines per centimeter); see Figure 8.

2. *Oscilloscope Application*

Operation of the Iatron storage tube probably can be understood best by noting how its characteristics apply to a particular application. Since the cathode-ray oscilloscope is commonly used

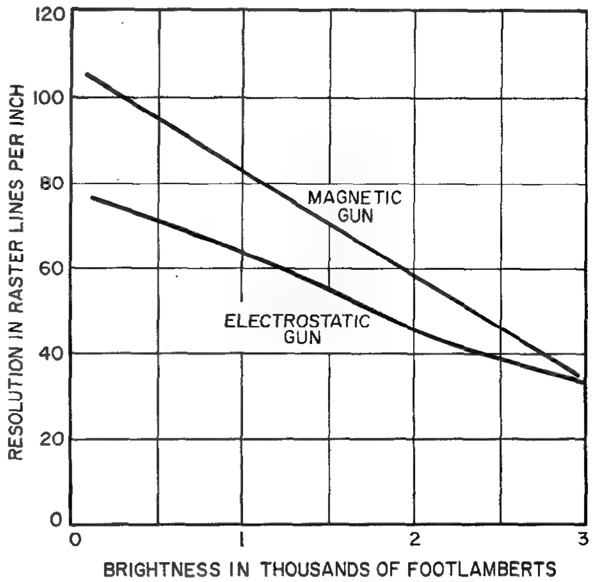


Figure 8—Resolution versus brightness. The curve for a magnetic-focus gun having a smaller writing-spot size shows limitations attributable to the writing gun.

in all engineering laboratories, this section will be concerned with the operation of the Iatron as it might be used in an oscilloscope.

The usefulness of oscilloscopes decreases rapidly for very-low sweep frequencies up to the threshold of flicker. The useful low-frequency range could be extended considerably by lengthening the persistence of the trace. Using the Iatron for this purpose, there is no flicker and the trace is bright enough to be viewed easily in a fully lighted room.

At low frequencies, storage in oscilloscope displays is ordinarily obtained by photographing the display. Another method to display low-frequency signals is to resort to mechanical means of recording. Besides the added expense and inconvenience of these methods, they also have limitations that are overcome by using the Iatron:

A. The Iatron will record transients composed of frequencies from direct current to above 1 megacycle, whereas mechanical recorders are limited to about 60 cycles.

B. Any trace can be stored for examination or it can be erased instantaneously in the Iatron.

C. The trace can be viewed immediately in the Iatron, avoiding the delay involved in development of film.

It is also practical to store superimposed sweeps taken in sequence at several test points for directly comparing waveforms.

The advantages of an Iatron oscilloscope are expected to be greatest at low sweep speeds, but it need not be restricted in operation to the low-frequency ranges, since a visible trace can be stored at writing-spot velocities up to nearly 10^6 centimeters per second.

At still-higher speeds, for which no trace will be stored, the tube can still be operated as a conventional cathode-ray tube, since the writing-beam average power input to the phosphor can be about 0.4 watt with the tube operating at 1 kilovolts; nor does the high voltage entail a loss of deflection sensitivity. A constant sensitivity of 100 volts per inch (39 volts per centimeter) is afforded by the electrostatic shielding property of the insulator screen, which isolates the 1-kilovolt deflection region of the tube completely from the 10-kilovolt phosphor potential.

By switching the insulator screen from its normal +10 volts to about −20 volts, the flooding beam can be cut off to improve contrast when it is desired to view only the writing-beam trace and not its stored image.

In normal operation, erasing pulses will keep the insulator erased to cutoff in areas where no trace is being written, and will prevent writing charges in the trace from integrating to the extent of charge spreading. For the usual repetitive-signal mode of oscilloscope operation, an erasing-pulse amplitude control and erasing-pulse width control should be accessible on the front panel to make initial adjustments of cutoff and viewing time. An intensity control is necessary to adjust writing-beam current to compensate for changes in sweep speed and waveform of the signal.

At sweep frequencies of over 45 cycles, the erasing pulse can be triggered by the sweep. This is in keeping with the discussion of writing-charge limiting in which it was pointed out that maximum charge stability of areas repeatedly written on exists when writing and erasing frequencies are equal. At lower sweep frequencies, flicker and blooming would be avoided if a constant erasing-pulse frequency of 45 cycles or higher were used. A convenient and satisfactory frequency is 60 cycles.

To display transients, maximum writing speed and storage time is desirable. A switch might be provided that, after writing, could be used to cut off the writing beam and erasing pulses simultaneously, thus avoiding over-writing of the transient trace and at the same time preventing its erasure. At extremely slow sweep speeds, it is desirable to turn off the erasing pulses before the start of the writing trace to avoid any erasure before one sweep is completed. This suggests a manual on–off erasing switch. Also, an instant-erasing button would probably be useful to restore the insulator quickly to cutoff after operating with the erasing pulse off.

The oscilloscope should be equipped with a z-axis gate to assure that the undeflected writing spot is cut off, since the undeflected spot would cause insulator charge spreading from that spot over an appreciable area of the screen and at

very-high current the insulator might even be damaged.

The controls described are the extent of the added complexity necessary to operate an oscilloscope adapted to the Iatron and the additional circuit needed to operate the flooding system is equivalent to adding one tube. An erasing-pulse generator that can perform the suggested functions could be a slave multivibrator.

Summarizing, the following controls are recommended for full utilization of the tube's capabilities:

A. Erasing-pulse gain control to adjust the amplitude of the pulses.

B. Erasing-pulse width control to adjust the duty cycle of the pulses.

C. Instant-erasing push-button switch that widens the erasing pulses momentarily to erase clutter without disturbing other erasing-control settings.

D. Erasing on off switch to remove erasing pulses when it is desired to freeze a trace for inspection without adjusting erasing controls.

E. Writing on–off switch to bias the writing-gun control grid to cutoff to prevent over-writing a frozen trace without adjusting the intensity control.

F. Flooding-beam on–off switch to bias the insulator support screen to flooding-beam cutoff when the tube is being used as a conventional cathode-ray tube.

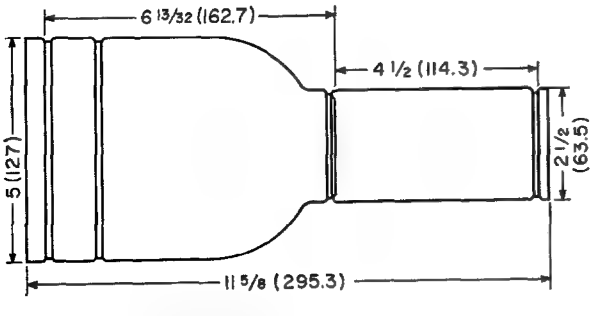


Figure 9—Outside dimensions of the 5-inch (127-millimeter) electric-field-deflection Iatron *Ia10P20-25*.

The type-*Ia10P20-25* Iatron shown in Figure 1 is the model recommended for oscilloscopes and other applications that require electric-field deflection. The useful display diameter is 4 inches (102 millimeters) and the outside dimensions are shown in Figure 9.

TABLE 1

Operating Voltages for Iatron*

Electrode	Voltage	Current
Writing Gun		
Heater	6.3	0.6 ampere, alternating or direct current
Cathode	−1000	1080 microamperes, maximum
Grid	−1042 at cutoff	
First Anode	−700 at focus	−1.0 microampere, maximum
Second Anode	40	940 microamperes, maximum
Flooding system		
Heater	2.5	2.5 amperes, alternating or direct current
Cathode	0	2.6 milliamperes
Anode and First Wall Electrode	40	0.8 milliampere, minimum 1.0 milliampere, maximum
Second Wall Electrode	20	0.07 milliampere minimum 0.115 milliampere, maximum
Third Wall Electrode	80	0.035 milliampere
Collector Screen	150	1.25 milliampere
Insulator Screen	+10	
Phosphor	+10 kilovolts, maximum	0.38 milliampere, maximum

* Deflection-plate reference voltage for minimum astigmatism, 0 volts. Deflection sensitivity; 85 volts per inch for plates D_3 to D_4; 100 volts per inch for plates D_1 to D_2. Plates D_1 and D_2 connected to +90 volts draw 36-milliampere flooding current.

Some comment on the operating circuit of Figure 2 is necessary. The resistance in the final deflection-plate circuit should be lower than is ordinarily used with cathode-ray tubes, because when they are driven positive, the plates can draw about 36 microamperes of flooding current.

If the average voltage of the deflection plates is about 40 volts, the least astigmatism of the writing spot results since the second anode and first wall electrode are at that potential. An astigmatism control consisting of a dual adjustable voltage divider could be inserted in the bleeder at the points supplying the direct current to the plates to adjust the average deflection plate voltage. However, it is found in practice that good results are achieved with an average plate voltage near zero, as shown.

The maximum phosphor voltage is 10 kilovolts. However, the characteristics of the tube, other than brightness, will be relatively unchanged with operation down to less than 5 kilovolts. Therefore, to avoid any possible damage to the tube because of overvoltage accidents, particularly when extremely high brightness is not an

objective, it is recommended that reduced voltage be used.

The flooding-spot size is adjusted by small changes in voltage applied to the second wall electrode after other voltages of the flooding system have been set at their specified values. Table 1 lists operating voltages and maximum and minimum currents of flooding-system electrodes that were measured to aid in the design of power supplies and bleeders. These measurements were made on only a few tubes, since production tubes were not available at this writing to obtain average data. It is anticipated that production tubes will have fewer electrodes, but those retained will be operated very closely to their present voltages and tubes will operate interchangeably, requiring only the number of controls that have been indicated.

Recent Telecommunication Development

Space-Charge Waves

RECENTLY PUBLISHED, a book on "Space-Charge Waves" and slow electromagnetic waves has been written by A. H. W. Beck of Standard Telecommunication laboratories. It is divided into 10 chapters, 12 appendixes, and sections on problems, references, and letter symbols.

Chapter 1—General Introduction

Chapter 2—Maxwell's Equations and Wave Equations

Chapter 3—Slow-Wave Structures

Chapter 4—Space-Charge-Wave Theory

Chapter 5—Matching Specified Input Conditions with Space-Charge Waves

Chapter 6—Space-Charge Waves in Klystrons

Chapter 7—Travelling-Wave Tubes and Backward-Wave Oscillators

Chapter 8—Crossed-Field Devices

Chapter 9—Special Space-Charge-Wave Devices

Chapter 10—Noise Phenomena in Space-Charge-Wave Devices

Appendix 1—Power Flow in a Circular Guide

Appendix 2—Integrals of Products of Bessel Functions

Appendix 3—The Tape Helix

Appendix 4—Variational Methods

Appendix 5—Measurements on Slow-Wave Structures

Appendix 6—The Focusing of Long Electron Beams

Appendix 7—Annular Beams and Bessel Function Expansions

Appendix 8—Solution of Equation (134) Chapter 4

Appendix 9—Orthogonal Expansions in Bessel Functions

Appendix 10—Coupling of Modes of Propagation

Appendix 11—Excitation of the Waves in T.W.A.s and B.W.O.s

Appendix 12—Llewellyn's Electronic Equations

This is the 8th volume in the International Series of Monographs on Electronics and Instrumentation published by Pergamon Press. The dimensions are $5\frac{3}{4}$ by $8\frac{3}{4}$ inches (14.6 by 22.2 centimeters). Of the 396 pages of text, the 10 chapters occupy 321 pages and include 124 figures and 916 numbered equations. The 72 pages of appendixes are predominently mathematical. There is a 3-page index.

The book is available from Pergamon Press, 4 Fitzroy Square, London, W1, at 90 shillings and from Pergamon Press, 122 East 55th Street, New York 22, New York, at $15.00 per copy.

Storage Tube Projects Radar Plan-Position-Indicator Display*

By HARRY W. GATES

Farnsworth Electronics Company, a division of International Telephone and Telegraph Corporation; Fort Wayne, Indiana

TO PROJECT live radar or beacon information directly onto a plotting board, two basic conditions must be fulfilled; high brightness and controllable long-time storage.

The radar projector indicator described in this article fulfills these requirements by utilizing a high-brightness storage tube called an Iatron®. The equipment, consisting of a range–azimuth projection indicator and a control console, is a remote plan-position indicator providing a 50-inch (1.27-meter) display for search radars.

To make the display equipment more useful, separate inputs for radar, rafax, and mapping are included. It is possible to select five radar ranges from 20 to 200 nautical miles (37 to 370 kilometers) and range markers from 5 to 50 nautical miles (9.2 to 92 kilometers). Sweep ranges are available for use with rafax at 120, 60, and 30 pulses per second. The indicator will accept radar triggers from 200 to 1200 pulses per second.

Antenna rotation information may be from 3 to 30 revolutions per minute and a two-speed synchro system insures accurate rotational information. A cycled instant erase automatically reduces build-up of ground clutter or large slow-moving rain-cloud formations.

1. *Storage Tube*

The basic elements of the Iatron storage tube are a writing beam of low intensity and high definition, a high-current flooding beam of large cross-section, a fine-mesh metallic screen supporting a thin insulating layer facing the electron guns, and an aluminized phosphor screen.

The writing beam scans the insulating layer and deposits charges everywhere proportional to the instantaneous beam intensity. Writing current is modulated by application of a video signal to the control grid. The distribution of potential over the insulating layer is then an electrostatic image of the video information.

Flooding-beam electrons approaching the charged insulating layer penetrate the fine-mesh screen in quantities proportional to the local charge on the insulator and impinge on the phosphor screen. In this way, the phosphor is continuously excited by the high-current flooding beam.

While writing and display are simultaneous, means are provided for insulator-charge erasure. If the insulator surface suddenly becomes more positive, flooding electrons strike the surface and charge it in the negative direction toward flooding-gun cathode potential. This is accomplished by applying a positive voltage pulse to the metallic screen supporting the insulator.

The most-negative potential to which the insulator will charge is determined by the erasing-pulse amplitude applied to the insulator support screen. The correct amplitude is that causing the equilibrium insulator potential after erasure to be equal to the potential required barely to cut off flooding-current flow to the phosphor. Then any insulator writing charge will be erased to cutoff after an adequate number of erase pulses of fixed width have been applied. Image persistence is controlled by adjusting pulse repetition rate.

2. *Servomechanism*

Many of the circuits employed in the indicator are conventional. However, several features are of interest.

The servo mechanism utilizes one-speed and 10-speed or one-speed and 36-speed selsyn information to synchronize the rotating sweep with the radar antenna. The one-speed system gives approximate information of the total train angle; the 36-speed system gives accurate indication of the angle but only if it is synchronized at the proper zero.

A switching system puts the one-speed system in control until the reproduced train angle is approximately correct and then throws the control to the 36-speed synchro system. Switching is accomplished by mixing the error signals in

* Reprinted from *Electronics*, volume 29, pages 172–175; December, 1956.

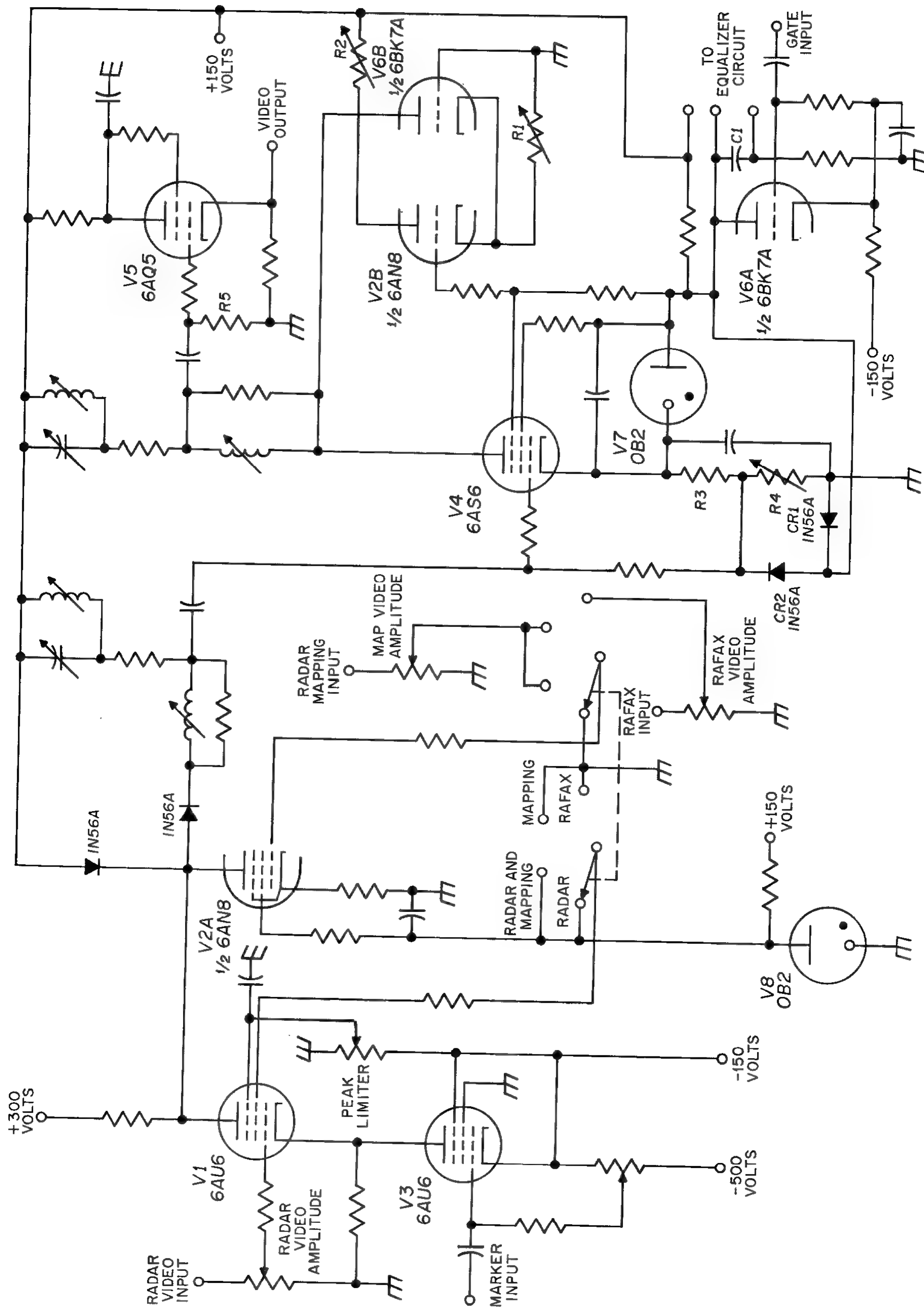


Figure 1—Radar, rafax, mapping, and marker signals are amplitude-controlled, mixed, peak-limited, and equalized in control console.

germanium diodes whose resistance decreases as the current flow increases. This insures a smooth switchover instead of a sudden switching such as is ordinarily encountered with synchronizing circuits using relays.

The video amplifier is composed of two sections. The first section, shown in Figure 1, is in the console where the radar video, mapping video, and marker signals are amplitude-controlled, mixed, peak-limited, and equalized (see also Figure 2). The second section, a two-stage amplifier, is in the projection unit (Figure 3) 25 feet (7.6 meters) away, which receives the video signal from the first section over a 100-ohm coaxial cable.

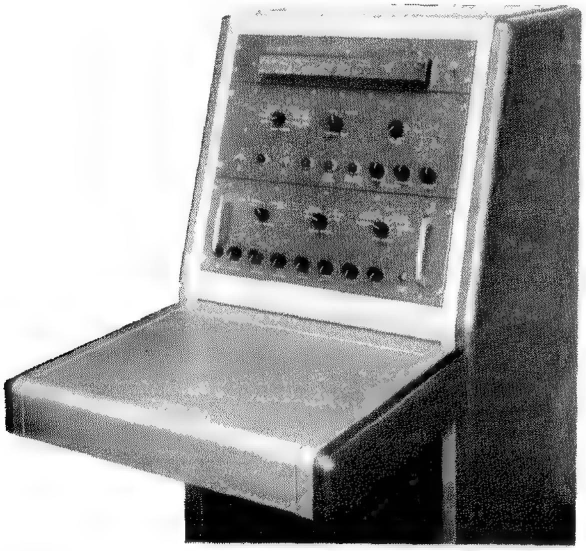


Figure 2—Control console for projector indicator.

3. Equalizing Circuit

Equalization of the video signal for a plan-position indicator presentation is accomplished in the second stage *V4*, of the video amplifier in Figure 1. With a rotating radial sweep, video signals of an amplitude that gives the proper brilliance near the periphery of the display tube would cause blooming near the center owing to the greater overlapping of adjacent scanning lines near the center.

To eliminate the blooming, less video gain is needed near the center of the display tube. This is accomplished by modulating the gain in accordance with a ramp function for a certain distance from the center and then continuing at constant gain for the remainder of the sweep.

The ramp-function waveform is applied to the suppressor grid of *V4*. Beside modulating the gain, the ramp-function signal of reverse polarity is also present in the plate circuit of *V4*. This is balanced out by the cathode-coupled amplifier consisting of *V2B* and *V6B*. The modulating ramp function is also applied to the grid of *V2B* which produces a signal of the same polarity in the plate circuits of *V2B* and *V4*. Adjustment of *R1* and *R2* will cancel the ramp-function signal in the plate load of *V4*.

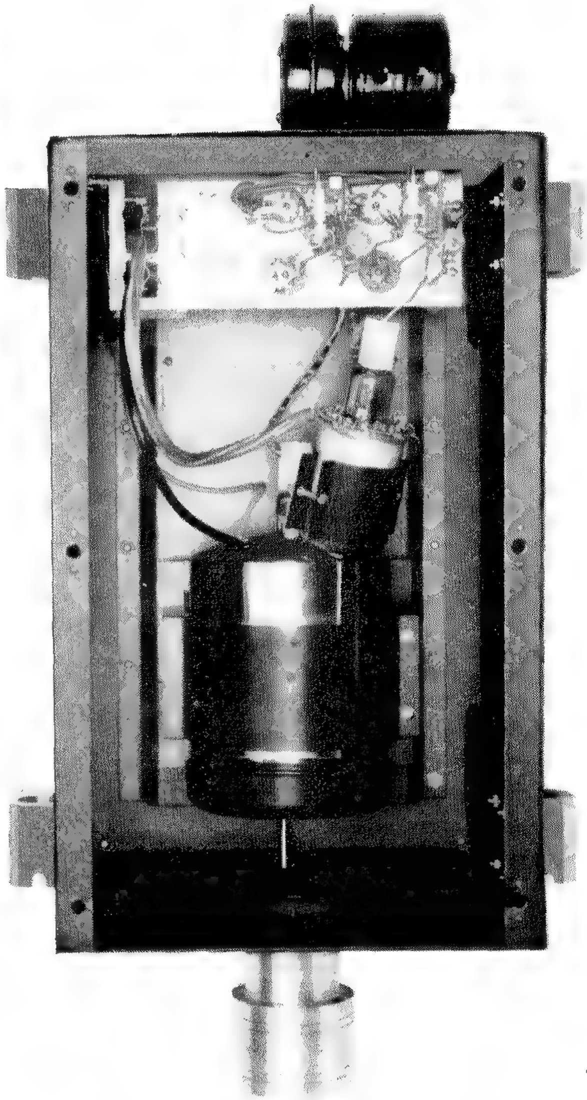


Figure 3—Inside of projector unit showing last stages of video amplifier and Iatron.

The ramp function is generated in the circuit associated with *V6A*. The cathode of *V6A* is at approximately −20 volts with respect to ground. Its plate is normally clamped to ground by *CR1*. The negative gate pulse, which initiates sweep, is applied to its grid.

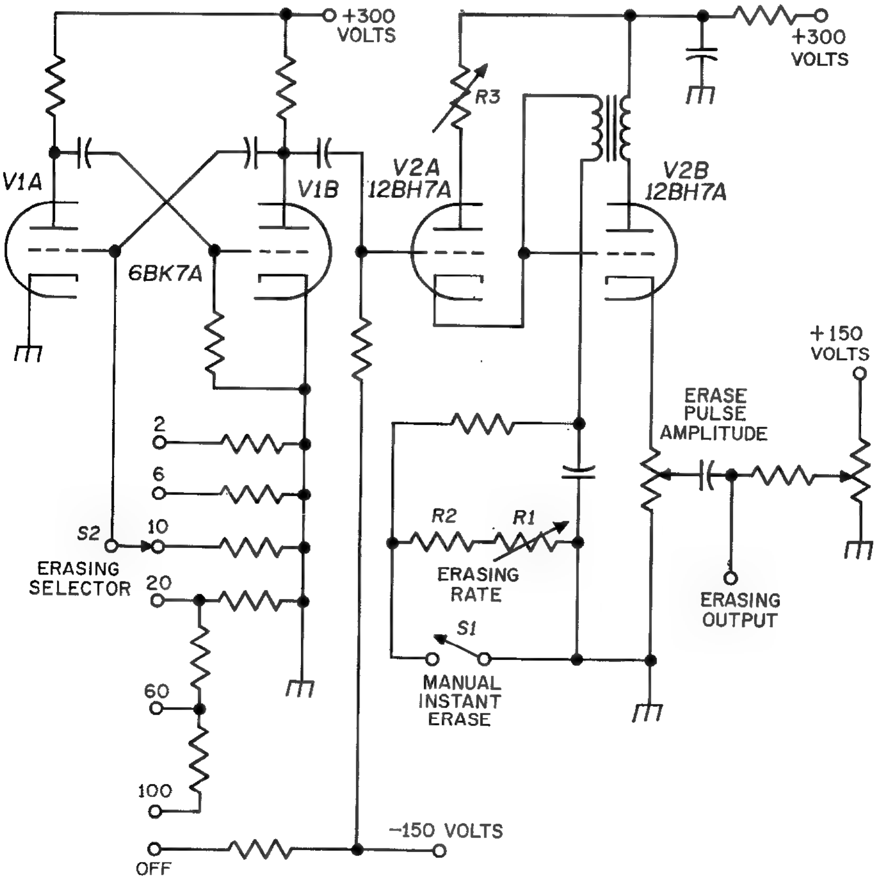


Figure 4—The pulse repetition frequency of the blocking oscillator controls the Iatron persistence.

Upon initiation of the sweep, the grid of *V6A* is driven negatively, cutting off the tube and permitting the plate voltage to rise as capacitor *C1* and a parallel capacitor, on another chassis, charge. When the plate voltage attains a value equal to that at the junction of *R3* and *R4*, *CR2* conducts, clamping the plate of *V6A* at this voltage, the maximum amplitude of the ramp function. For the remainder of the sweep it remains constant at this value.

The equalizing circuit chassis contains a capacitor and adjustable charging resistance for each range, with a separate adjustable charging resistance for each of the expanded ranges. Screwdriver controls permit adjustment of the ratio of the time the ramp function reaches its maximum to sweep time for each range and each expanded range.

Referring to Figure 1, the positive video signal from *V4* is fed to the grid of *V5*, a *6AQ5* used as a cathode-follower to change the impedance for coupling into a 93-ohm coaxial cable connecting with the projection unit. The 100-ohm cathode load resistor for *V5* is at the output end of the cable in the projection unit. Resistor *R5* prevents the cathode of *V5* from floating if the cable is disconnected while the equipment is turned on.

4. *Erasing Generator*

The Iatron persistence is controlled by varying the pulse repetition frequency of a blocking oscillator. The erasing-rate control on the console enables the operator to select any continuously variable persistence from 3 milliseconds to 20 seconds.

The erasing generator is a blocking oscillator, *V2B*, with the output taken from the cathode circuit as shown in Figure 4. Erasing-rate control *R1* controls the blocking-oscillator frequency. Pushbutton switch *S1*, when depressed, shorts *R1* and *R2*, causing the blocking oscillator to operate at a high frequency, which instantly erases the insulator screen.

Automatic instant erasure is provided at fixed intervals by stable multivibrator *V1*. The grid circuit of *V1A* has a long time constant that can

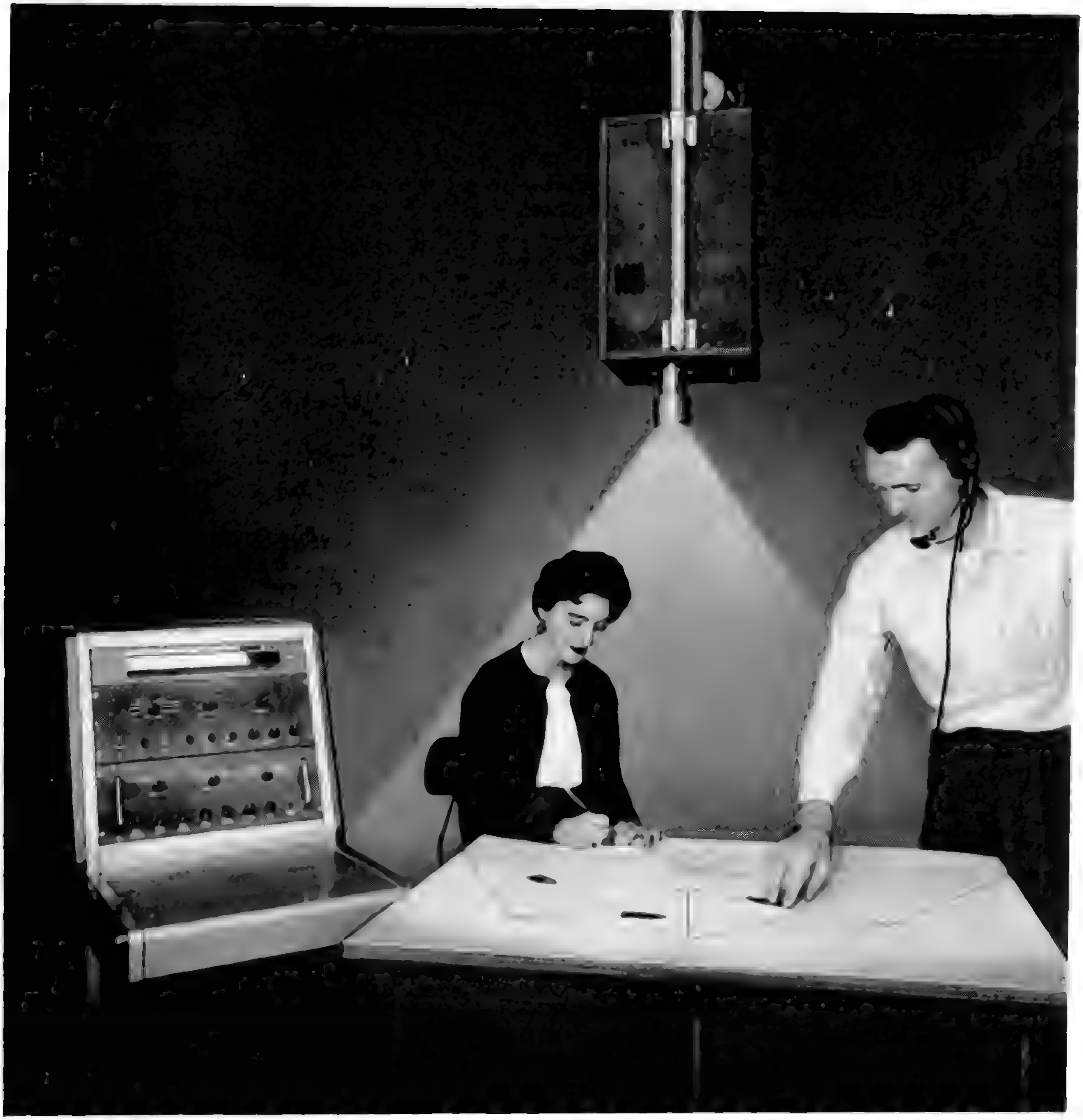


Figure 5—Radar projector indicator as used with plotting board for air-traffic control.

be varied by $S2$, causing the period to be varied in steps from 2 to 100 seconds.

The grid of $V1B$ has a relatively short time constant. When the plate of $V1B$ goes positive, the grid of cathode-follower $V2B$ goes positive, causing the grid of the blocking oscillator to go positive. This prevents the blocking-oscillator grid from being driven negative due to the cathode-follower, which causes it to operate at a high frequency for the relatively short period that the plate of $V1B$ is positive. Rheostat $R3$ controls the output impedance of the cathode-follower and hence the instant-erasure frequency. When the cathode-follower has a high impedance, the grid of the blocking oscillator can be driven slightly negative, decreasing the frequency.

5. *Application*

The projection unit of the indicator is normally mounted at the ceiling of the control room and the information displayed on a horizontal plotting board (Figure 5). However, the projection unit can be mounted in any position for vertical projection on a screen or projection from below through a translucent surface.

Simple refractive optics eliminate the problems of adjustment encountered in the more-complex Schmidt-type optics. A single knob is the only adjustment.

When the indicator is used in conjunction with the rafax video-bandwidth compression system for remote radar or beacon information, a novel and interesting display is produced. Such a system, illustrated in Figure 6, was demonstrated by the Technical Development Center of the Civil Aeronautics Administration in Indianapolis.

Aircraft equipped with beacon equipment were interrogated by the beacon station in Jamestown, Ohio. The beacon video, trigger, and rotational information were transmitted by microwave link to Dayton, Ohio. At Dayton, the beacon information was reduced in bandwidth by the rafax encoder. This reduced-bandwidth information was then sent over a 100-statute-mile (161-kilometer) telephone line to Indianapolis where it was displayed by the Iatron radar projection indicator on a plotting board.

Another indicator simultaneously projected beacon information from the Indianapolis beacon on the same plotting board. The composite display was then used for traffic-control evaluation for the surrounding 400-nautical-mile (761-kilometer) radius.

The storage time of the Iatron was set for slightly greater than one antenna revolution so pertinent information was displayed at all times. Aircraft were identified and shrimp boats placed on the plotting board to be moved along as the flights progressed.

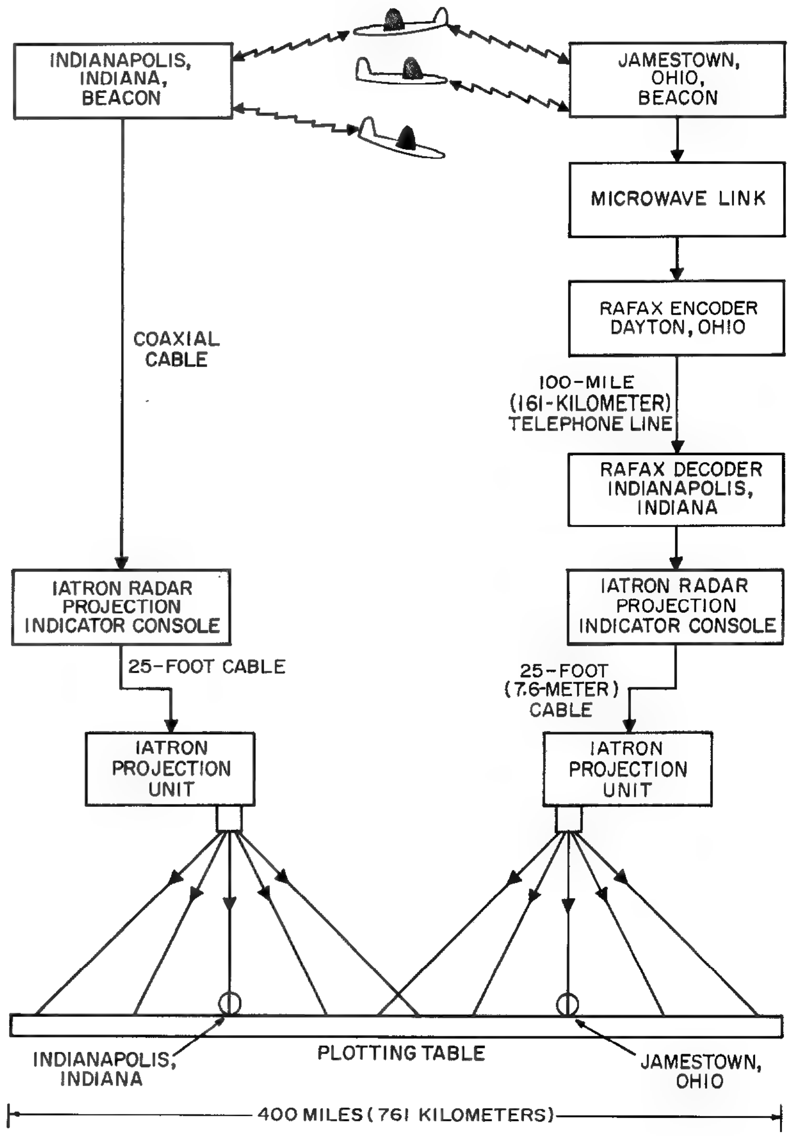


Figure 6—Civil Aeronautics Administration traffic control system using radar projector indicator.

Carrier Telecommunications Over the High-Tension Power Network in Algeria

By EDMOND A. THOMAS

Compagnie Générale de Constructions Téléphoniques; Paris, France

BEFORE the second world war, several companies had initiated the construction of networks for distributing electricity in Algeria, but their development remained limited by the comparatively low industrial activity of the country. After the war, it became apparent the increases in electric-power consumption would raise a serious problem and that a large interconnection network would have to be constructed. A working plan was then drawn up by the electricity division of the Gouvernement Général de l'Algerie to be developed by a newly formed organization, Electricité & Gaz D'Algerie (EGA).

Five years were required to erect a vast distribution network that received its main supply from 4 steam and 9 large hydroelectric plants.

With an annual production of the order of 800 million kilowatt-hours, Algeria is now in a position to meet its future needs, including those resulting from the electrification of the railways.

The large dams that are part of the hydroelectric plants were also planned to improve irrigation, which is always critical in North Africa. This important work is one of the French undertakings that honors those who were responsible for its planning and execution.

1. *High-Tension Network*

The more-densely populated region of Algeria, a strip of cultivatable land bordering the Mediterranean and stretching back to the edge of the desert sands, covers an area about 1100 kilometers (684 miles) long and 100 kilometers (62 miles) wide. The high-tension network necessarily took the same form. To connect the large towns, which are mainly scattered along the coastline, a main trunk line operating at 150 kilovolts was established to connect the 90- and 60-kilovolt distribution networks. This is shown in Figure 1. As rapid means of communication were not available, it was decided that an intercommunication telephone system would have to be superposed on the power network.

In 1949, a contract was signed for the application of carrier telephony to the existing and planned 90- and 60-kilovolt sections and to the 150-kilovolt principal trunk line then under construction. Automatic switching was to be provided for the whole network and would employ 3-digit dialing.

2. *High-Frequency Carrier System*

The first part of this program involved the installation of more than 20 carrier telephone installations with average lengths of about 100 kilometers (62 miles). The equipment had to be simple, economical, and easy to maintain. A standardized amplitude-modulated type that has been used on many power lines in metropolitan France proved to be completely suitable when manufactured for tropical service.

The transmitter includes a crystal-controlled oscillator followed by a two-stage amplifier. A carrier power of 15 watts is developed for continuous sine-wave operation. The modulation chain includes an amplifier that is adjusted so that at the zero level of 1 milliwatt on the telephone channel, the corresponding modulation of the carrier is about 35 percent.

The receiver employs two untuned radio-frequency stages preceded by a high-quality filter having a pass band of 3 kilocycles per second each side of the carrier. It provides attenuation of at least 45 decibels to interference from a carrier 6 kilocycles away.

This type of transmitter–receiver is used for duplex transmissions in the 50- to 300-kilocycle range, allowing a minimum space of 25 kilocycles between the transmitting and receiving frequencies. The selection of these two frequencies is made by associated filters at the common connection point to the power line.

On most of the connections, the high-frequency transmission is also employed for protecting the

line. For this purpose, the telephone channel is subdivided by filters that select the 300- to 2300-cycle band for speech. If a fault occurs on the line, speech is replaced by a 2580-cycle signal sent at a minimum modulation of 80 percent. This signal reduces the tripping time of the breaker at the far end of the section.

Three of these equipments are shown at the right in Figure 2.

The second part of the program was more difficult because on the long 150-kilovolt trunk line the dispatching center at Algiers had to be connected with the maximum of speed and reliability to the secondary dispatching centers at Bône and Oran and to the entire system. Long distances are involved and the line voltage was high enough to require more-complex equipment to obtain a satisfactory signal-to-noise ratio.

For these long lines, single-sideband equipments have been installed in groups of three to provide parallel channels on adjacent frequencies. Only two of these channels are used under normal conditions. The third is a spare that can be substituted for either of the other two in case of failure. If needed, it can be used for the transmission of telemetering data. The changeover of equipments can be effected quickly by switches in a connecting cabinet, which can be seen in the center of Figure 2. Connected to the lower part of it is a cabinet for connecting the three channels in parallel. A differential system employs a coaxial cable to connect the group of three bays to the single input device to the power line.

The single-sideband equipments shown at the left of Figure 2 are mounted on removable chassis in apparatus bays. Each chassis has a definite function and contains vacuum tubes and a group of air-tight boxes. They are connected to the general wiring of the bay by means of lateral rows of plugs, which allow numerous tests to be carried out.

The transmitter is designed for a theoretical bandwidth of 4 kilocycles. The frequencies transmitted are obtained by two successive modulations. The first is under control of a 24-kilocycle crystal oscillator. The upper sideband from 24 to 28 kilocycles is selected and converted to the operating frequency range between 50 and 300 kilocycles through the second modulation.

The crystal frequencies for this second modulation are spaced at 4-kilocycle intervals, placing side by side the transmission channels of the bays functioning in parallel in the same transmission direction.

The output amplifier for each bay can deliver a sine-wave power of 30 watts, but as a result of the loss introduced by the differential system,

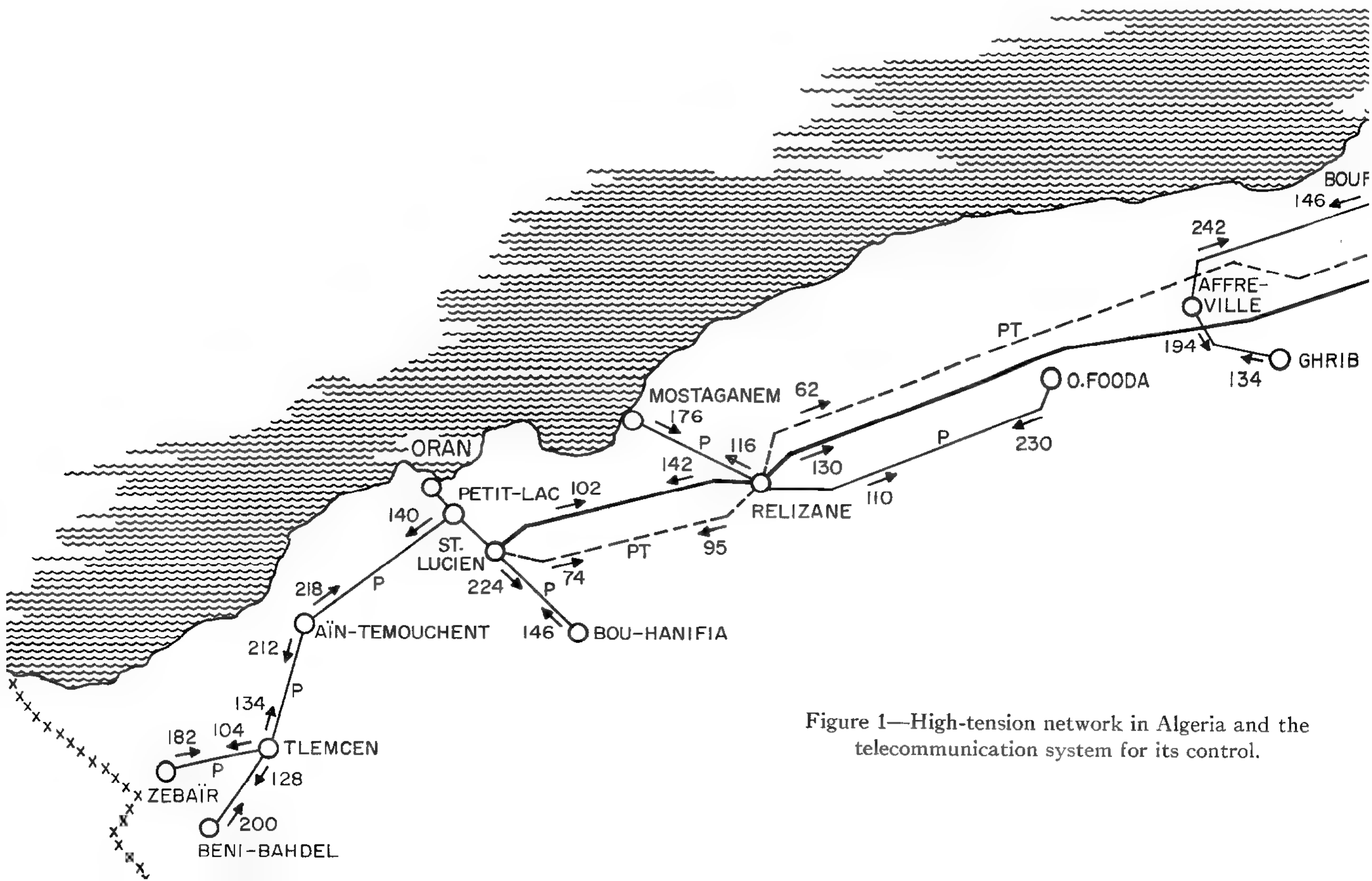


Figure 1—High-tension network in Algeria and the telecommunication system for its control.

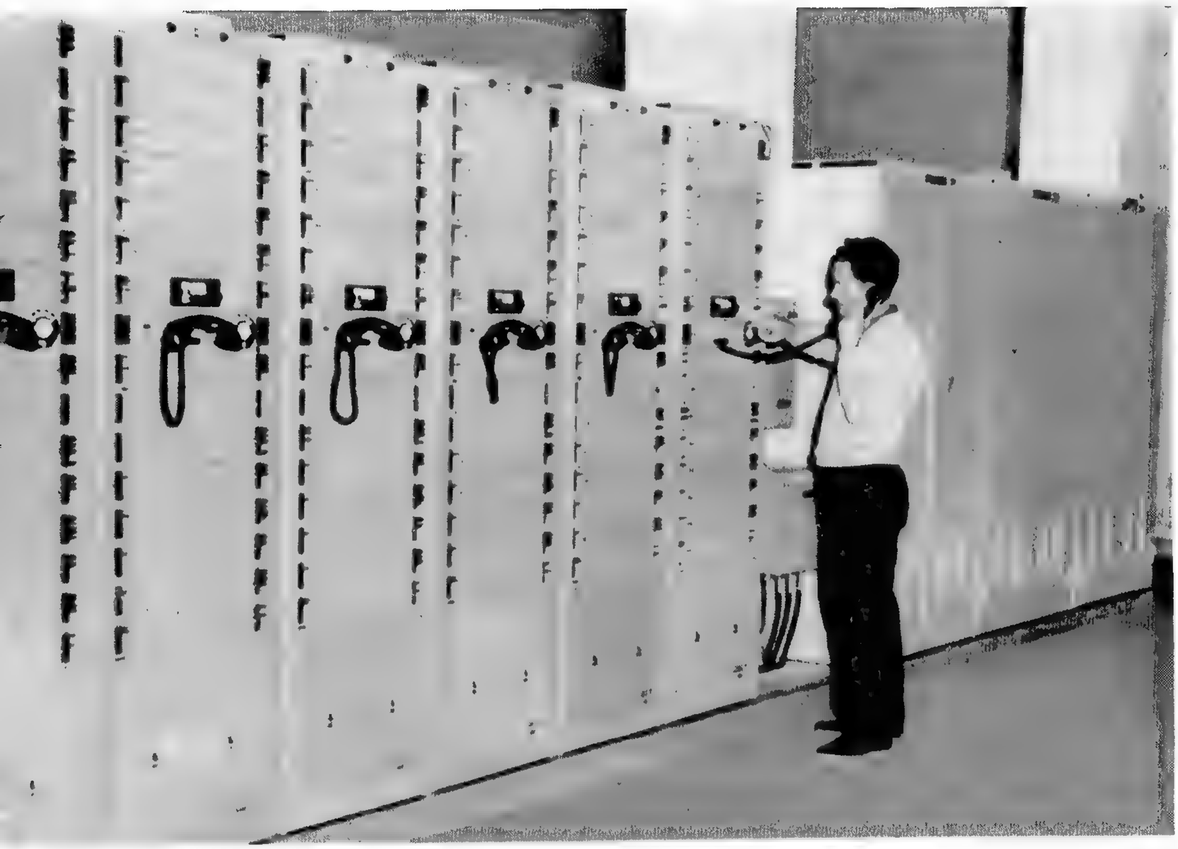
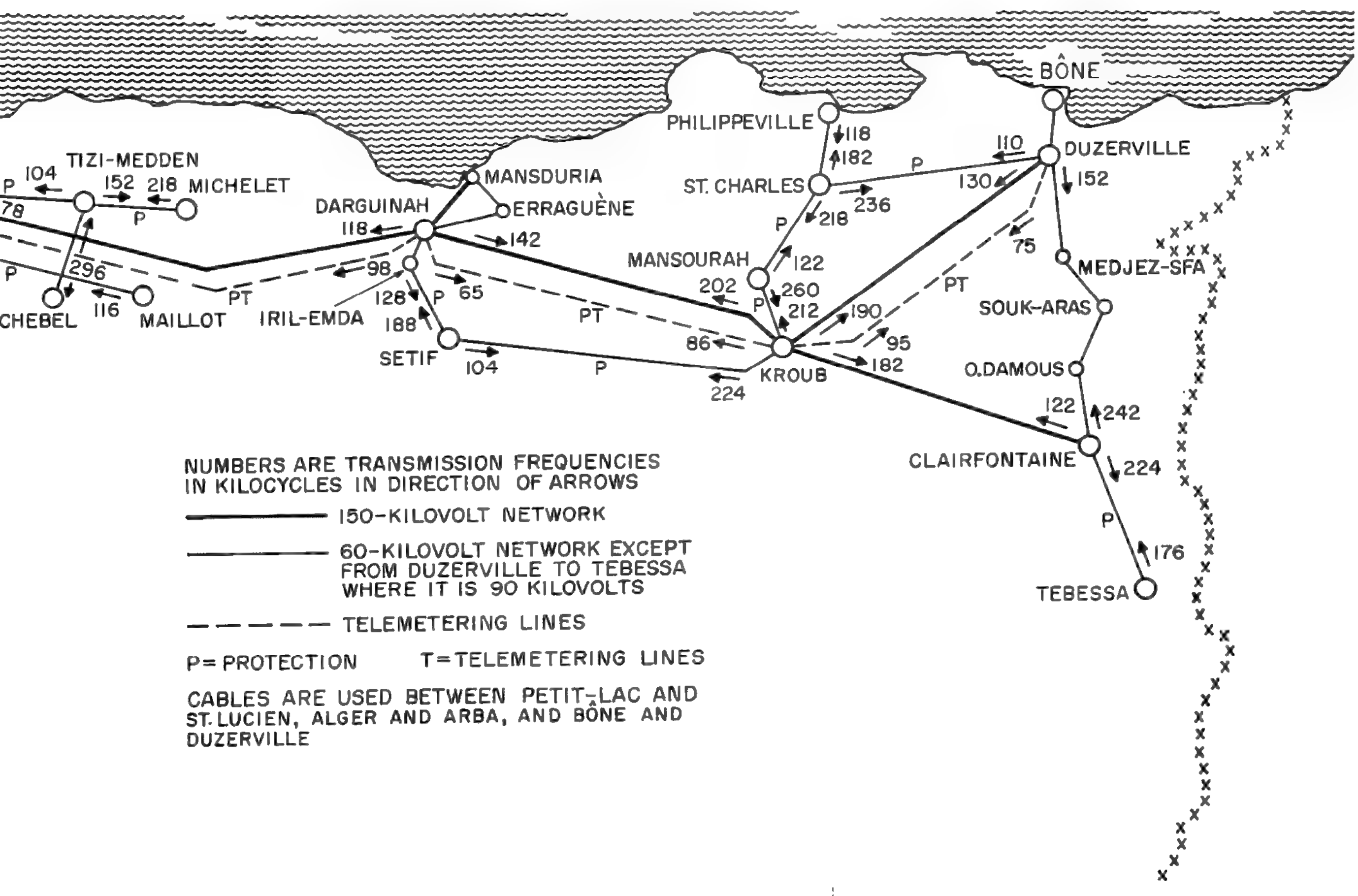


Figure 2 –High-frequency equipment at Arba. The single-sideband equipment is at the left and the double-sideband apparatus is at the right. In the center are the changeover switches for the three channels.



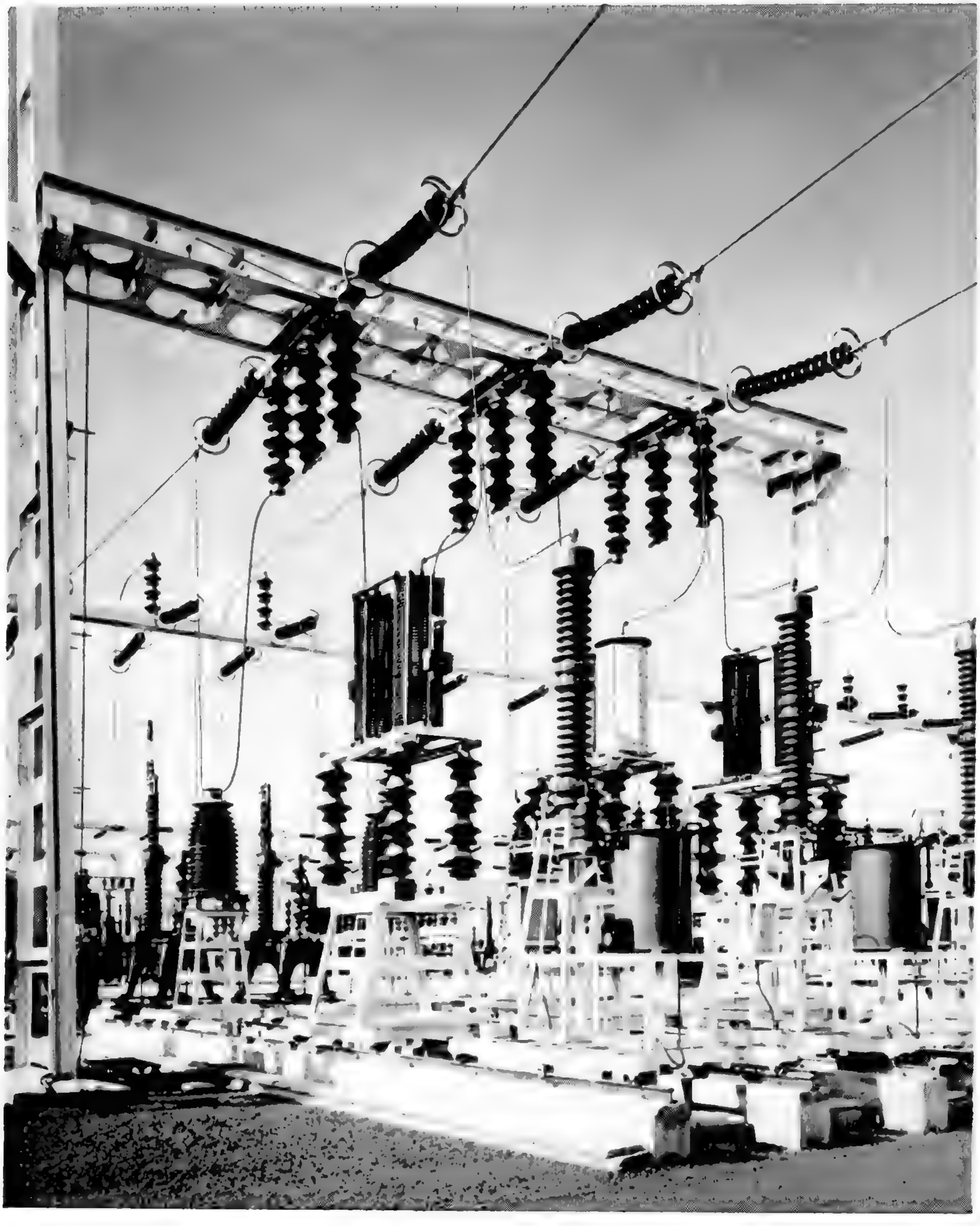


Figure 3—Line coupling installation at Duzerville. The petticoated device in the center, to which is connected the downlead from the line phase nearest the camera, is the coupling capacitor. At the base of the coupling capacitor are the drain coil and the protective devices and, partially hidden behind the framework, the line-tuning cabinet. The large vertical inductors are carrier-current line traps. The cylindrical tank contains part of the 150-kilovolt capacitive voltage divider.

the corresponding power delivered to the line is only of 10 watts in the case of three bays.

The first element in the receiver is a band-pass filter that selects the frequencies before the first demodulation. This demodulation, which is controlled by a crystal identical with that of the corresponding transmitter, brings the received frequencies back into the 24- to 28-kilocycle band. A later filter having very-sharp cutoff characteristics provides the interchannel selectivity. The voice frequencies are restored exactly to their original values by the second demodulation, which uses the 24-kilocycle voltage transmitted in an attenuated form and amplified in the receiver. This attenuated pilot is, moreover, used for the gain control since it changes with line attenuation. Regulation is obtained by controlling the heating of thermistors incorporated in an attenuator.

The theoretical 4-kilocycle channel is subdivided so as to provide a band from 300 to 2300 cycles for telephony. A channel of 2580 to 3060 cycles can thus be reserved for transmission of five frequencies in the overtone series (odd multiples of 60 cycles). Dialing is obtained through interruption of a 3300-cycle wave, the highest frequency used in the band allotted to each channel.

Of the 30 bays of this type required for the 150-kilovolt trunk line, 20 are in permanent operation for telephone traffic, which can be repeatered 4 times for communication between the most-distant dispatching centers of Oran and Bône.

Among the three dispatching centers, one of the two high-frequency channels is reserved for priority communications and the other is used for ordinary communications.

3. *Line Equipment*

At all substations, the high-frequency equipment is connected to the line input channeling equipment by an underground armored coaxial cable. This cable enters the coupling box, which contains a transformer with taps for adapting it to the line impedance, and is also connected to a tuning device. The tuning system generally includes the capacitor providing coupling to the power line as part of a high-pass filter. These capacitors are moreover used as voltage dividers connected to voltage transformers; at their base they are equipped with a discharge coil to bypass dangerous voltages, and with the usual protective devices, such as lightning arrestors and fuses. This subassembly appears in the center of Figure 3. At the side of the capacitors of the extreme phases may be seen the tuned circuits provided on the station side for blocking the high-frequency currents of an interphase link.

4. *Telephone System*

The third part of the program consisted of the organization of the telephone system. The various carrier connections to the 150-, 90-, and 60-kilovolt networks were considered to be equivalent to 4-wire telephone channels terminating in 2-wire circuits through differential devices. However, in the case of long-distance communication, the retransmissions between consecutive sections are carried out on 4 wires; the equipments are then used as base-to-carrier-frequency repeaters. It should be added that the communication channels are extended from the 150-kilovolt line to the switching stations at Algiers, Oran, and Bône through multichannel cables about 30 kilometers (19 miles) long, in accordance with the standards of the Comité Consultatif International Télégraphique et Téléphonique.

To provide automatic switching over the whole system, each subscriber's set is identified by a 3-digit number. The system is divided into 7 districts, to each of which is assigned a different figure in the hundreds position. The tens represent the various stations, and the units positions are allotted to the individual subscriber's sets in each station.

All the switching operations are effected by automatic switches, the capacity of which varies according to the size of the station and the amount of traffic. These switches are derived from four basic types that are standardized for use in similar systems in France. Racks of these switches are shown in Figure 4. They have the following capacities.

Type *C* has a capacity of 8 carrier circuits between power switching stations and 10 subscribers' lines.

Figure 4—Racks of telecommunication switches at Algiers and terminations of cables to Arba.

Type *B* has a capacity of 14 carrier circuits between power switching stations and 20 subscribers' lines.

Trada type *1* accomodates 1 trunk or main line and 6 subscribers' lines.

Trada type *2* accomodates 6 trunks or main lines and 40 subscribers' lines.

It should be added that to allow for the special connections of the carrier channels on the 150-kilovolt trunk line, two types of automatic switches are provided. Priority communications are controlled by special automatic switches that allow only priority communications between dispatching centers and between dispatching centers and certain subscribers' sets in the network. Ordinary switches normally serve all the subscribers' sets in the network and can, through their connection with the other switches, come to the aid of the priority subscribers if their network should be overloaded.

As the power network is paralleled in places with the 90- and 60-kilovolt lines, the automatic switches have been designed so that they can select a free path if the called subscriber can be reached over several routes.

Finally, as a security measure, all the large 150-kilovolt stations are equipped with a desk that allows priority telephone connections to be made manually in case of failure of the local automatic switching system. One of these desks is shown in Figure 5.

5. Conclusion

This system, which has been made entirely automatic through the installation of specially designed telephone switches, provides, by means of 3-digit dialing, for quick and reliable intercommunication over the whole territory with the standards of quality specified by the Comité Consultatif International Télégraphique et Téléphonique.

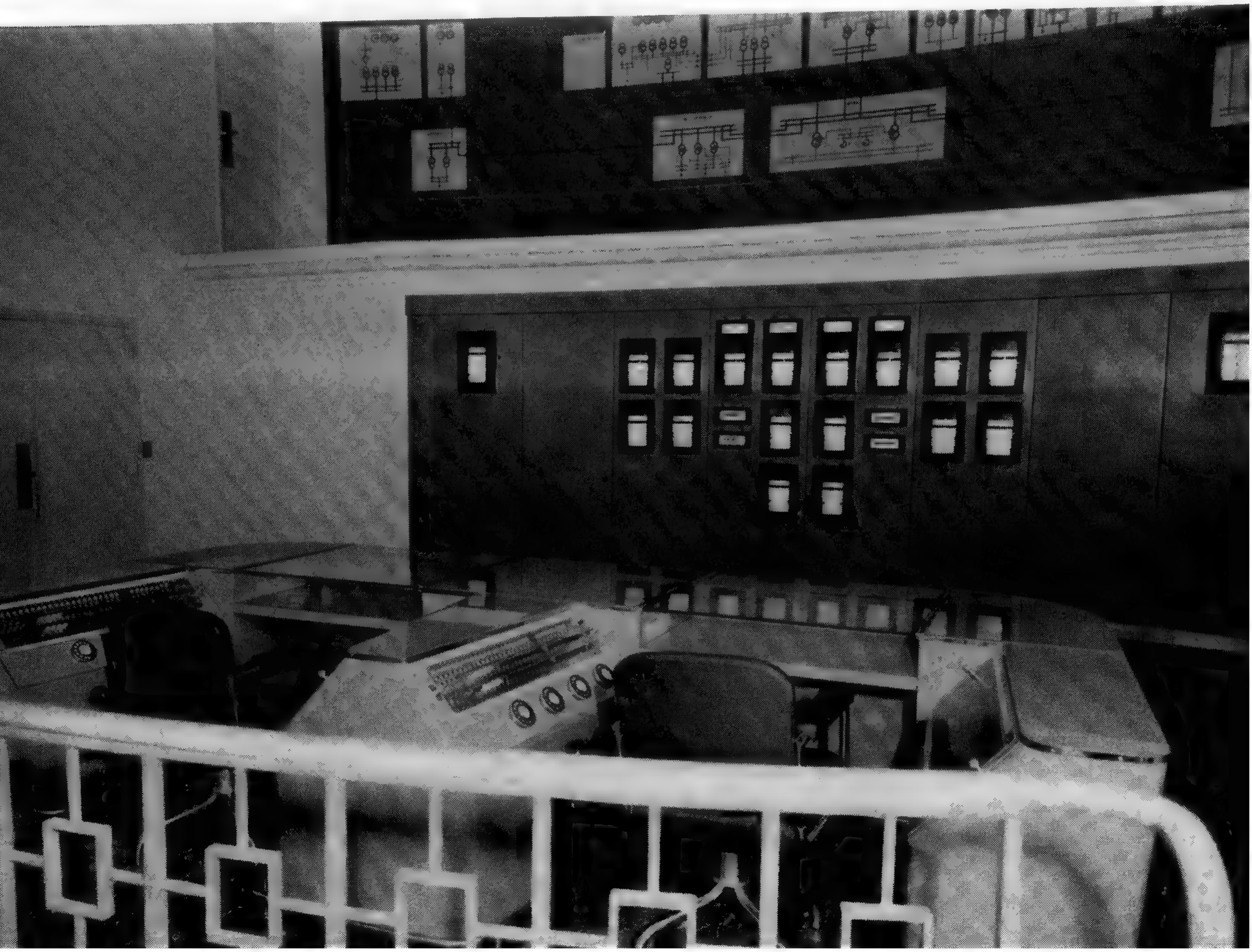
The network will be further expanded through interconnection with the Moroccan network now under construction. It allows the control centers at Algiers and Casablanca to supervise the distribution and exchange of power.

6. Acknowledgment

Compagnie Générale de Constructions Téléphoniques considers it a privilege to have collaborated with the Equipment Construction Division of Electricité & Gaz D'Algerie in providing the Algerian power distribution network with an autonomous telecommunication system.

Much information and the photographs in this article were kindly supplied by Mr. Grau of that organization.

Figure 5—Regional power-dispatching desk at Algiers. This was supplied as part of the telecommunication installation.



Crossbar Toll Offices in the Finnish Group Network*

By NIKOLAUS LEWEN and HEINZ KÜRTEN

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CROSSBAR automatic telephone exchanges used as local offices with a maximum capacity of 100 lines in the subzone areas of the Finnish network are discussed. The trunking principles, testing facilities, and constructional features are described.

• • •

1. *Finnish Telephone Network*

The country is subdivided into 80 groups of which one is mapped in Figure 1. The toll traffic between these groups is switched through 9 toll distribution exchanges. Each group exchange is connected to at least one of the toll exchanges; direct lines between group exchanges are also provided and can alternatively be switched to the toll route. The toll exchanges are completely meshed. It is intended to introduce a subscriber toll dialing system whereby the first three digits dialed, of which the first will be a *9*, will put the subscriber through to any of the group exchanges. (Helsinki will be reached by dialing *90*.) During interim operation, toll calls will be established by dialing *09*, which will connect the calling subscriber to the toll operator's position in his group exchange.

The entire toll traffic between toll exchanges is in the hands of the public administration, while terminal exchanges are operated either by private or communal telephone companies. For the sake of uniform operation, it is intended to combine various local organizations within a group area.

A linked numbering scheme with a maximum of five digits containing the code of the terminal exchange is used within each group. Six-digit numbers are provided only for the Helsinki area.

The terminal exchanges of a group are radially connected to the group exchange via tandem exchanges. The latter are linked by direct lines, which are also provided between tandem exchanges of adjoining groups. The idea is at present being considered of establishing direct lines even between terminal exchanges of the same and of adjoining groups, if necessary.

Owing to the various proprietors of terminal exchanges, a number of different systems are in use and plants of different manufacture are encountered within one group. The administration of the national system, in its status as a supervisory authority, has issued recommendations and instructions aiming at more uniformity, at least of newly acquired switching plant.

2. *Crossbar Terminal Exchange*

In accordance with conditions in a thinly populated country, it is intended to equip terminal exchanges with crossbar switches serving primarily up to 50 and, with expansion, up to 100 subscribers. Each crossbar switch frame contains facilities for 50 subscribers.

Exchanges of this size have no permanent maintenance attendants. A functional check is carried out at intervals, mostly on meter readings.

Crossbar switches[1] and magnetic counters[2] are, due to their high reliability, particularly suitable for unattended operation.[3]

To check the conditions prevailing in a system from an attended central maintenance point, a so-called automatic subscriber has been provided in every terminal exchange. By dialing a certain number, this automatic subscriber will report by tone signals on the operational condition of the plant. In addition, an alarm sending device in the terminal exchange will signal disturbances that may occur in any of the more-important units of the switching plant. These signals are received in the tandem exchange.

Every line engagement causes the terminal exchange to be connected by a junction to the

* Reprinted from *SEG-Nachrichten*, volume 3, number 1, pages 35–38; 1955.

[1] J. Bernutz, "Der Koordinatenschalter KS 53," *SEG-Nachrichten*, volume 2, number 1, pages 6–10; 1954.

[2] R. Scheidig, "Der Zählmagnet ZM 53," *SEG-Nachrichten*, volume 2, number 1, pages 11–13; 1954.

[3] A. Mehlis, "Wähler oder Schalter als Verbindungsorgane in der Fernsprechvermittlungstechnik," *Fernmeldtechnische Zeitschrift*, volume 5, pages 293–296; 1952.

andem exchange, where a central register–ranslator switches into the circuit, takes all lialing digits, and remains connected to the ine until it receives the end-of-selection signal.

The dialed zone, as established by the register–translator, is passed on to a metering pulse sender permanently connected to the junction; as soon as the conversation begins, metering

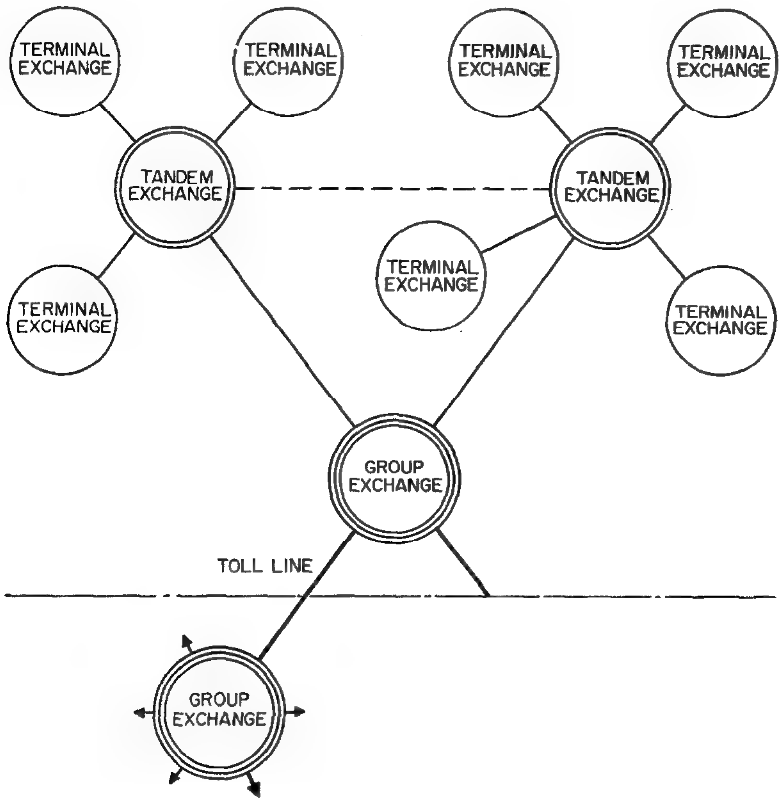

Figure 1—Layout of a group within the Finnish network.

This register–translator performs the following operations.

A. Route control; that is, selection of the shortest possible route (direct or tie lines between tandem exchange).

B. Determination of dialed zone for message metering.

C. Prevention of false engagements beyond the tandem office.

D. Upon storage of long-distance digit *9*, connection of a long-distance storage device in the group exchange to route the call.

pulses are transmitted to the calling subscriber. The interval between pulses corresponds to the zone and decreases with increasing distance.

All signals are transmitted through a junction as alternating-current signals and converted to direct-current signals at the receiving end. Since the relay repeaters are designed for simultaneous transmitting and receiving, only two signals, differentiated by their length and sequence, are necessary. This simplifies the engineering and reduces lost time. The repeater relays in any intermediate exchanges operate as pulse repeaters for a connection, requiring no counting devices. Hence, the code scheme can be modified without modification of the relay repeaters.

Using additional plant, terminal exchanges

can operate with carrier systems as well, both with in-band and out-of-band signaling, the latter being the case in which signals are regenerated in the carrier system.

The busy signal, or revertive blocking, can be a continuous signal in lines without amplifiers. In lines with amplifiers, blocking is brought about by pulsed signals to avoid overloading the group amplifiers.

the subscriber's last two digits, one magnetic counter KS to control sequence of dialing, and one magnetic counter KM checking the exchange code, which is part of the five-digit directory number. All dialing information for long-distance and short-haul (internal) traffic is received and evaluated by these magnetic counters. When a call remains within the terminal exchange, an internal linking circuit is engaged to check,

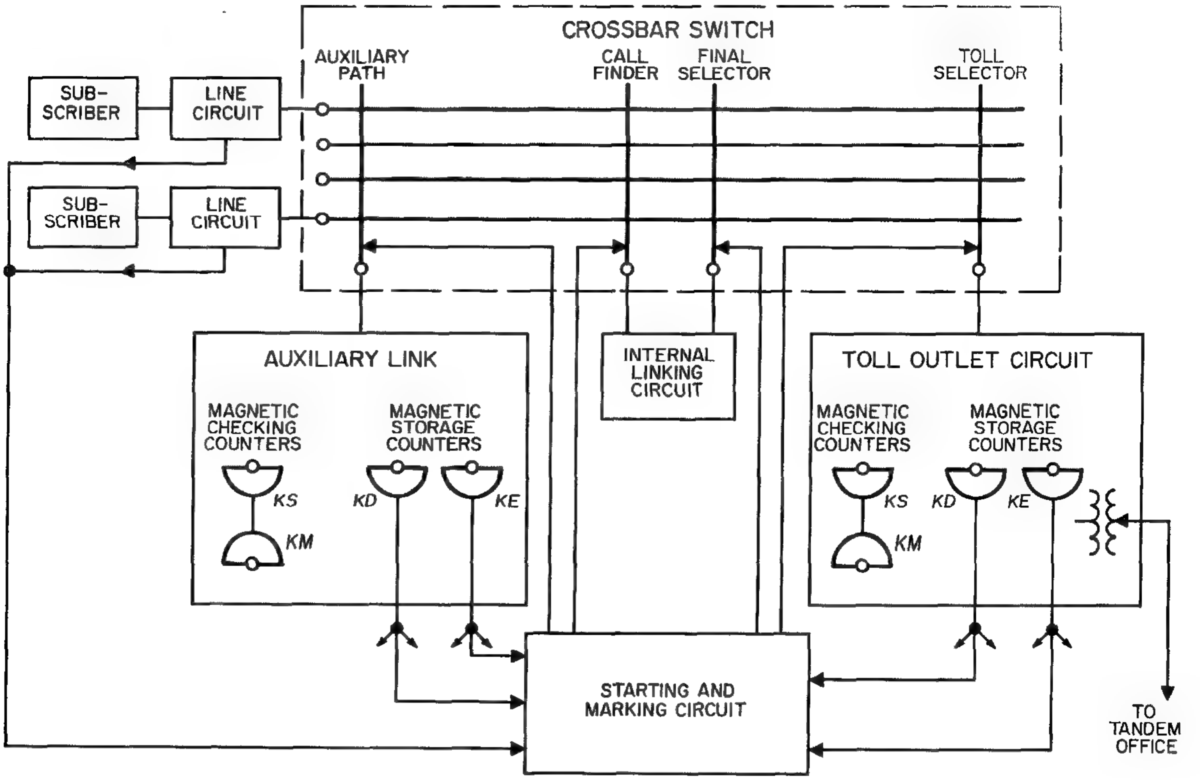

Figure 2—Starting and marking of junctions in a crossbar exchange.

3. *Establishment of Connections*

The establishment of a connection will be explained with reference to Figure 2. The junctions between terminal and tandem exchanges are bidirectional. In the terminal exchange, this junction leads to the long-distance outlet, connecting to one of the vertical bars of the crossbar switch. In the incoming direction, this vertical bar operates as a trunk final selector; in the outgoing direction as a call finder. Each external linking circuit has two magnetic counters, KD and KE, for the storage of the call, connect lines, and meter. Accordingly, this internal linking circuit is comparatively simple.

3.1 External Connections

For an external connection or a long-distance call, the subscriber actuates the starting and marking circuit by lifting the handset. These circuits connect a free trunk selector to the subscriber's line. At the same time, the tandem exchange end of the junction is coupled to a register–translator. The subscriber's dial pulses

are stored by the register in the tandem exchange and by the magnetic checking counter in the terminal exchange. Since the called subscriber is connected to an external terminal exchange, the dialed exchange code will not correspond to the code of the calling subscriber's exchange. In this case, the subscriber's terminal exchange does not accept further dial pulses. The evaluation of the pulse series for further routing is accomplished in the register of the tandem exchange alone.

The first revertive pulse reaching the calling subscriber's terminal exchange is the starting signal. It actuates the first counting pulse applied to the subscriber's meter and, for coin boxes, the reversal of the line conductors. Other subsequent counting pulses are emitted by the pulse sender in the tandem exchange and are counted during the conversation.

3.2 Internal Connection

When a subscriber of the same terminal exchange is being called, the long-distance outlet is seized in the same way as in the case of an external call. However, the office code dialed is now that of the same exchange; as soon as the magnetic checking counter has noted this identity, a clear-forward pulse will prevent any false engagement of the same terminal exchange on a second junction. When the last two pulse series have adjusted magnetic counter *KD* and *KE* for storage, the long-distance outlet circuit will connect to the marking circuit via a common lockout chain. The marking circuit has two functions.

A. To connect an internal linking call finder to the calling subscriber's line.

B. To connect the final selector of this internal link with that subscriber's line marked by magnetic storage counters *KD* and *KE*.

When the connection between calling and called subscribers is established—which takes about 150 to 200 milliseconds—the long-distance outlet releases the marking circuit and terminates the pulse, clearing the junction to the tandem exchange. The internal linking circuit can meter the conversation either by one pulse per conversation or by pulses timing the length of the conversation, depending on local conditions. In the first case the metering pulse is emitted at the end of conversation.

If the called subscriber is busy or if no free path in the internal linking circuit is available, all circuits engaged by the calling subscriber are immediately released and a busy tone is transmitted to the subscriber. If all long-distance outlets are busy when a subscriber tries to establish an internal connection, this connection can still be established via an auxiliary link. This auxiliary link will also release all circuits and transmit the busy tone to the calling subscriber in the case where the called subscriber belongs to another terminal exchange.

3.3 Incoming Traffic

In the case of an incoming call, the exchange code has been absorbed by the preceding selector stages; the two remaining digits actuate *KD* and *KE* in the same way as for local traffic. The called line is correspondingly marked and the crossbar-switch vertical bar is switched. The long-distance outlet circuit gives the signals for end of dialing, called-subscriber answering, and clearing. Where an incoming call is connected by an operator, there is the possibility of the operator's cutting into an existing conversation.

4. Ancillary Plant

Two-party lines, coin boxes, and grouped-number service (private-branch-exchange line hunting) are provided. The various numbers of a grouped number are not bound to a certain sequence, which would confine them to the same decade. Without any switching procedure, they can be used as night call numbers. In a fully equipped terminal exchange, eleven lines are additionally available; for instance, for grouped numbers or for coin boxes.

5. Testing Equipment

Tests of all switching facilities and metering circuits are provided. The switching plant can also be tested separately.

5.1 Testing Long-Distance Outlet Circuits, Outgoing Direction

When the test relay set (Figure 3) is connected to jack *J2* of the long-distance outlet circuit, the junction to the tandem exchange is revertively blocked. The cord of tester *T1* is connected to jack *J1*. Brief depression of the ground button in *T1* engages the long-distance outlet and generates the engaged pulse. The *KA1* relay in the test transmission circuit receives the engaged pulse, the subsequent dial pulses, and the clear-forward pulse, which are then evaluated by a test setup or made visible by an indicator lamp.

The revertive called-subscriber answering and metering pulses are initiated by a short depression of the counting key; replacing the handset is followed by a release guard signal pulse.

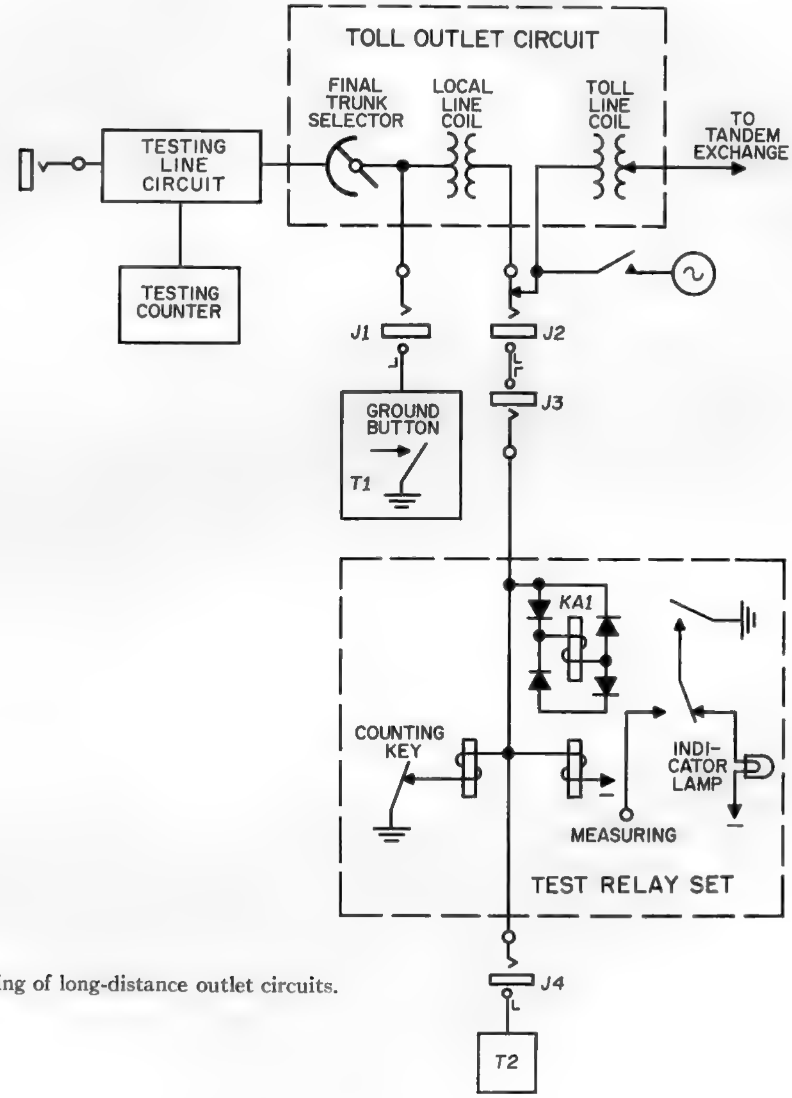

Figure 3—Testing of long-distance outlet circuits.

5.2 Testing Long-Distance Outlet Circuits, Incoming Direction

The number of the automatic subscriber serves also as a test number. The automatic subscriber is connected through jack contacts and is disconnected by the engagement of this jack with tester *T2*.

When digit *1* is dialed, *T1*—now connected to jack *J4* to establish the connection—transmits

an engaged pulse. Now the number of the automatic subscriber is dialed; the latter is connected to *T2*. In this procedure, revertive pulses are evaluated by the test relay circuit.

Instead of calling *T2*, any one of the ordinary subscribers' lines may be called.

5.3 Testing Internal Link Circuits

Tester *T1* is connected to the jack of the internal link circuit under test while *T2* is again connected to the jack of the automatic subscriber (Figure 4).

When engaged by depressing the button, the

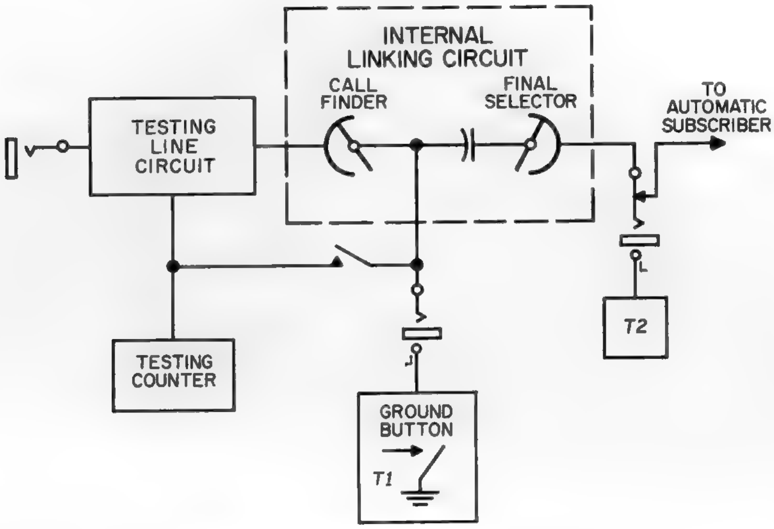

Figure 4—Testing of internal linking circuits.

internal linking circuit automatically adjusts itself to the directory number of the automatic subscriber and performs all functions including metering.

5.4 Operational Test

This test requires connection of the testing line circuit to *T1*. It permits establishment of all connections outgoing from the crossbar-switch terminal exchange. The subscribers' lines are connected to the terminal exchange via a disconnecting distributor. When the plugs are disconnected, the circuits can be tested either with or without the subscribers' lines.

6. *Trunking Scheme*

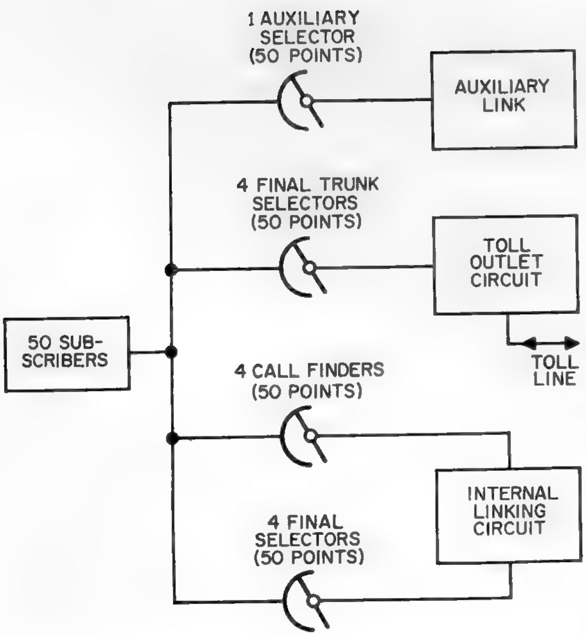

Figure 5—Trunking scheme for 50-subscriber equipment.

As will be seen from Figure 5, the crossbar terminal exchange is equipped with circuits for 50 subscribers, 4 long-distance outlets, 4 internal linking circuits, and 1 auxiliary link. When

equipped to full capacity (100 subscribers), the terminal office comprises 6 long-distance, 6 internal, and 1 auxiliary circuits. The long-

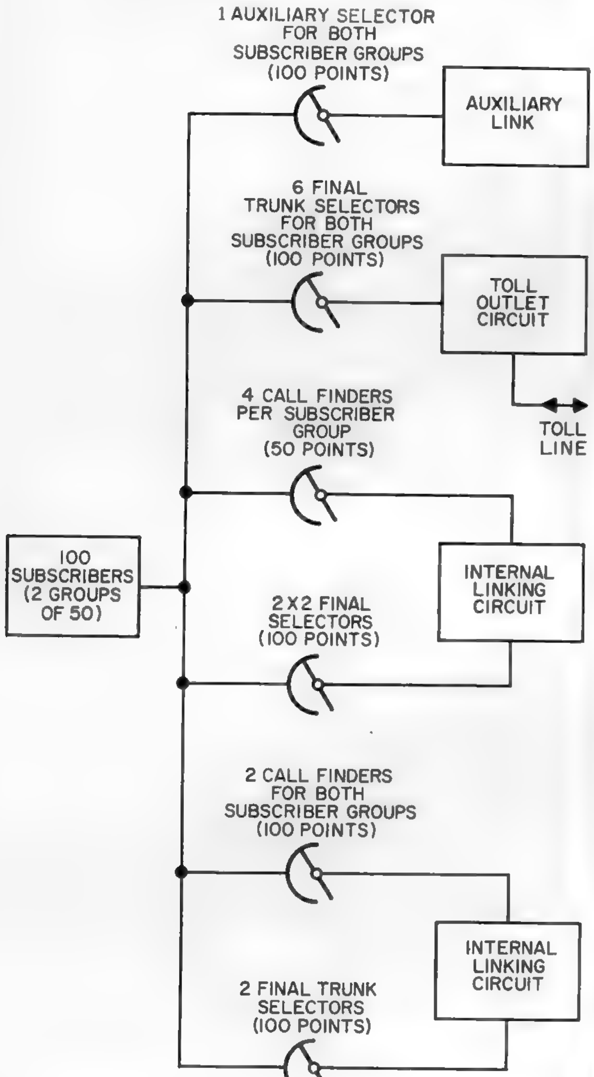

Figure 6—Trunking scheme for 100-subscriber equipment.

distance outlets, the auxiliary link, and the final selectors have a total of 100 points. The call finders of the internal linking circuits are subdivided into 4 call finders with 50 points or outlets handling the normal traffic and two 100-point call finders for the peak traffic. This is shown in Figure 6.

7. Construction

The crossbar exchange is mounted in a bay about 7 feet (2.1 meters) high, 40 inches (1 meter) wide, and about 20 inches (0.5 meter) deep, as shown in Figure 7. All switching circuits are arranged on two hinged frames and

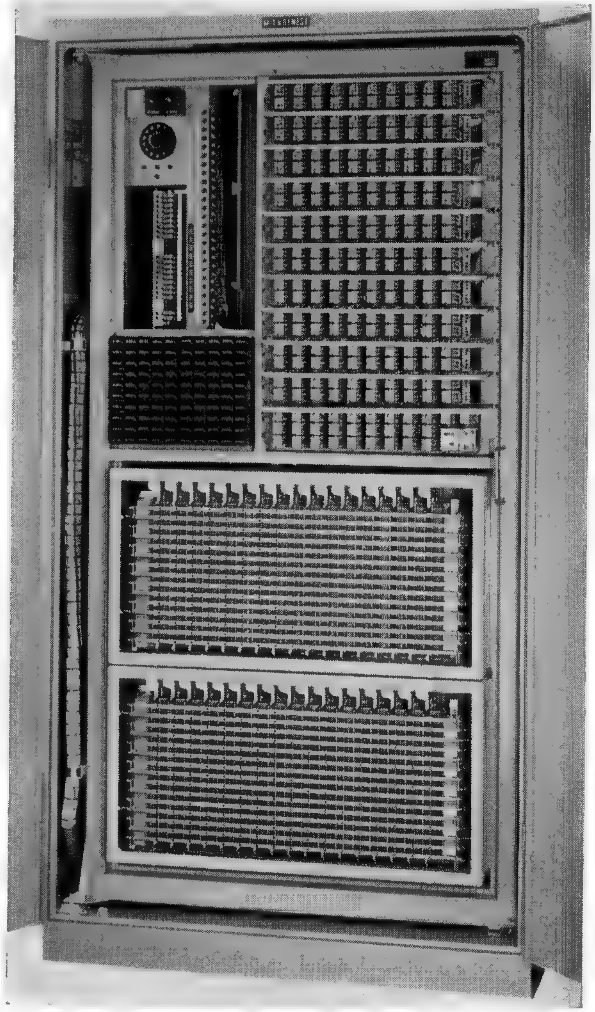

Figure 7—Crossbar terminal exchange.

are easily accessible. The front frame comprises the subscribers' relays, metering circuits, the crossbar switch, and the central marking circuit. The rear frame carries the relays of the long-distance outlet circuits, internal links and auxiliary link, as well as signaling and testing equipment, jacks, and push buttons.

Two-wire Repeater Employing Negative Impedance*

By THEODOR G. GREWE

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COMPENSATION for losses in a nonloaded cable can be accomplished with a simple circuit having attenuation and impedance values that are similar in the voice-frequency band to those of a cable for which the circuit may be substituted. The desired properties are obtained with a bridge network consisting of a balancing inductance and two simple resistance–capacitance branches. These branches are converted into negative impedance by two single-stage transistor amplifiers. The result is an equivalent circuit having, through its negative impedances, the properties of a two-wire repeater.

1. *Application*

The repeater is designed for voice-frequency-operated, nonloaded two-wire lines. The circuit components determining the characteristic impedance and the gain can be inserted as individual units, which makes the repeater readily adaptable to cables having various conductor diameters. Particular care was exercised to obtain a characteristic impedance similar to that of standard cables. The remaining net loss of a repeater section is small and is only slightly dependent on the location of the repeater. The gain can be adjusted so that the net loss becomes independent of frequency. For instance, with a cable loss of 10 decibels at 800 cycles per second and 19 decibels at 3400 cycles, a reduction in attenuation to about 4 decibels can be achieved.

2. *Reduction of Attenuation by Negative Resistances*

Assume a structurally nonsymmetrical T network consisting of the resistances $\mathbf{R}_1$, $\mathbf{R}_2$, and $\mathbf{R}_3$ as shown in Figure 1A. The attenuation of this network can be compensated by connecting to the series resistances $\mathbf{R}_1$ and $\mathbf{R}_2$ two equivalent negative resistances $-\mathbf{R}_1$ and $-\mathbf{R}_2$ and in parallel with the shunt resistance $\mathbf{R}_3$ a negative resistance $-\mathbf{R}_3$ having the same absolute value. This is shown in Figure 1B.

The same effect can be obtained by connecting the three negative resistances as a separate T network. Alternatively, both T networks may be combined into a π network so that, at the junction points, structural symmetry of opposing signs is formed. This is shown in Figure 1C. Resistances $\mathbf{R}_3$ and $-\mathbf{R}_3$ are connected without loss by the resistances $\mathbf{R}_2$ and $-\mathbf{R}_2$. The value of the parallel resistances $\mathbf{R}_3$ and $-\mathbf{R}_3$ is infinite, and the resistances $\mathbf{R}_1$ and $-\mathbf{R}_1$ cancel each other.

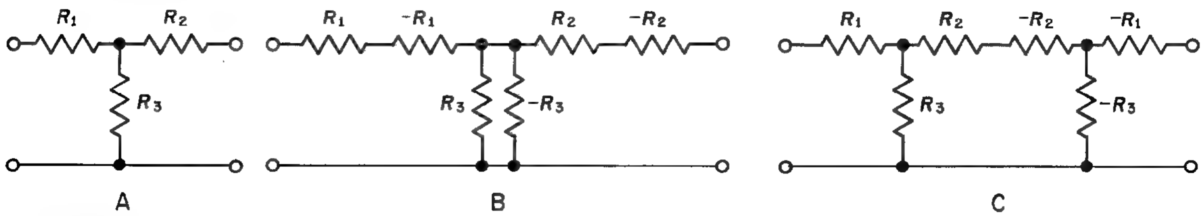


Figure 1—The attenuation of the nonsymmetrical T network at *A* may be compensated for by the insertion of corresponding negative impedances at *B*, which may also be accomplished at *C* by converting to an equivalent π network.

Within a limited frequency band, a telephone line can be approximated by a structurally symmetrical T network. Hence, the attenuation of this line could be cancelled by a symmetrical T network consisting of three negative resistances. However, the use of an equivalent circuit consisting of a bridge network with a symmetrical balancing inductor is advantageous because the number of required negative resistances is reduced to two.

* Originally published under the title "Ein Zweidrahtverstärker mit negativen Widerständen" in *Nachrichtentechnische Zeitschrift*, Volume 8, pages 610–618; November, 1955.

3. *Properties of Bridge Network*

According to Bartlett's theorem,[1] there is an equivalent bridge network for any symmetrical four-pole circuit. In the case where such a circuit represents a line of length $2l$ (see Figure 2A), the equivalent circuit is a lattice network shown in Figure 2B, where the two series impedances of the line legs $\mathbf{Z}_s$ must be equal to the short-circuit resistance and the two shunt impedances $\mathbf{Z}_o$, must be equal to the open-circuit resistance of a cable having the length l.

In the frequently used equivalent bridge network having a balancing inductor in one leg (see Figure 2C), the series impedance must be of twice the value of the short-circuit resistance and the shunt impedance must equal half the value of the open-circuit resistance. The same applies to another bridge network where the short-circuit and the open-circuit values are merely interchanged as in Figure 2D. These resistances refer to the cable length l.

The ideal negative line for compensating the attenuation of a cable of length $2l$ could, then, be constructed as shown in Figure 3. The passive impedances balancing the open-circuit and short-circuit impedances are changed by converters into negative impedances. Due to the finite bandwidth of the converter components, however, this conversion can be achieved only approximately and for a limited frequency band. The balancing of the cable impedance can also be only approximated by a limited number of components. Therefore, the properties of a bridge network having two arbitrary impedances $\mathbf{Z}_1$ and

[1] R. Feldtkeller, "Einführung in die Vierpoltheorie der Nachrichtentechnik," Hirzel, Leipzig; 1948.

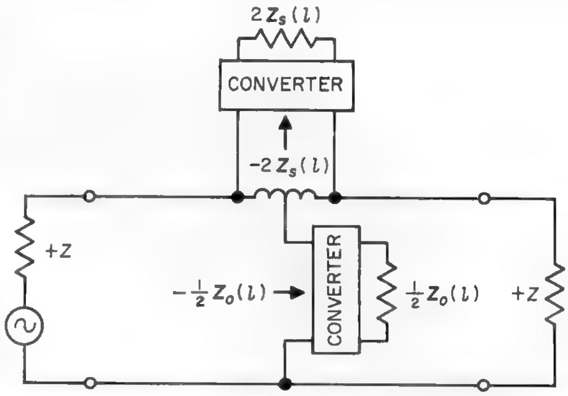


Figure 3—Construction of ideal negative line to compensate for the attenuation of a cable of length $2l$.

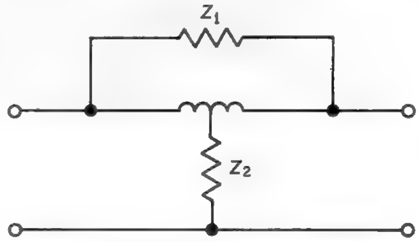


Figure 4—Bridge network having two arbitrary impedances.

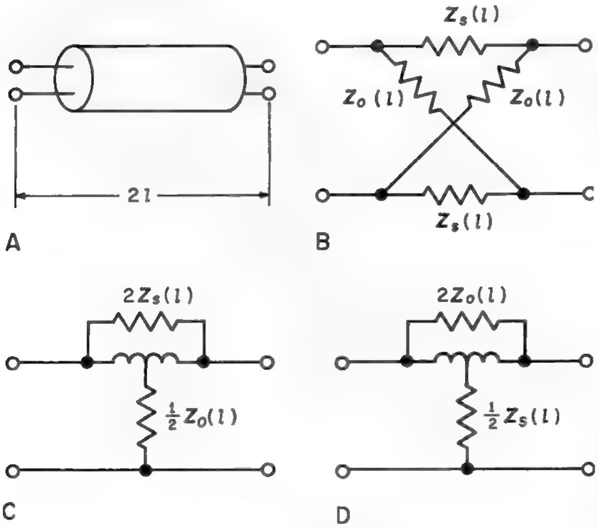


Figure 2—Equivalent lattice or bridge networks for a two-wire line.

$\mathbf{Z}_2$ will be discussed first. For the characteristic impedance $\mathbf{Z}$ and the image transfer constant $g = a + jb$ of the circuit shown in Figure 4, the following two equations are valid.

$$\mathbf{Z} = (\mathbf{Z}_1\mathbf{Z}_2)^{1/2} \tag{1}$$

$$g = a + jb$$

$$= \ln \frac{1 + \frac{1}{2}(\mathbf{Z}_1/\mathbf{Z}_2)^{1/2}}{1 - \frac{1}{2}(\mathbf{Z}_1/\mathbf{Z}_2)^{1/2}}. \tag{2}$$

Since the characteristic impedance depends only on the product and the image transfer constant depends on the ratio of the impedances $\mathbf{Z}_1$ and $\mathbf{Z}_2$, the values for $\mathbf{Z}$ and g can be prescribed independently of each other. If the bridge-network impedance $\mathbf{Z}_1$ is related to the characteristic impedance by $\mathbf{N} = \mathbf{Z}_1/\mathbf{Z}$, where $\mathbf{N}$ is generally

the complex function of the frequency ω, then using (1),

$$\mathbf{Z}_1 = \mathbf{NZ} \tag{3}$$

$$\mathbf{Z}_2 = \mathbf{Z}/\mathbf{N} \tag{4}$$

and (2) can be replaced by

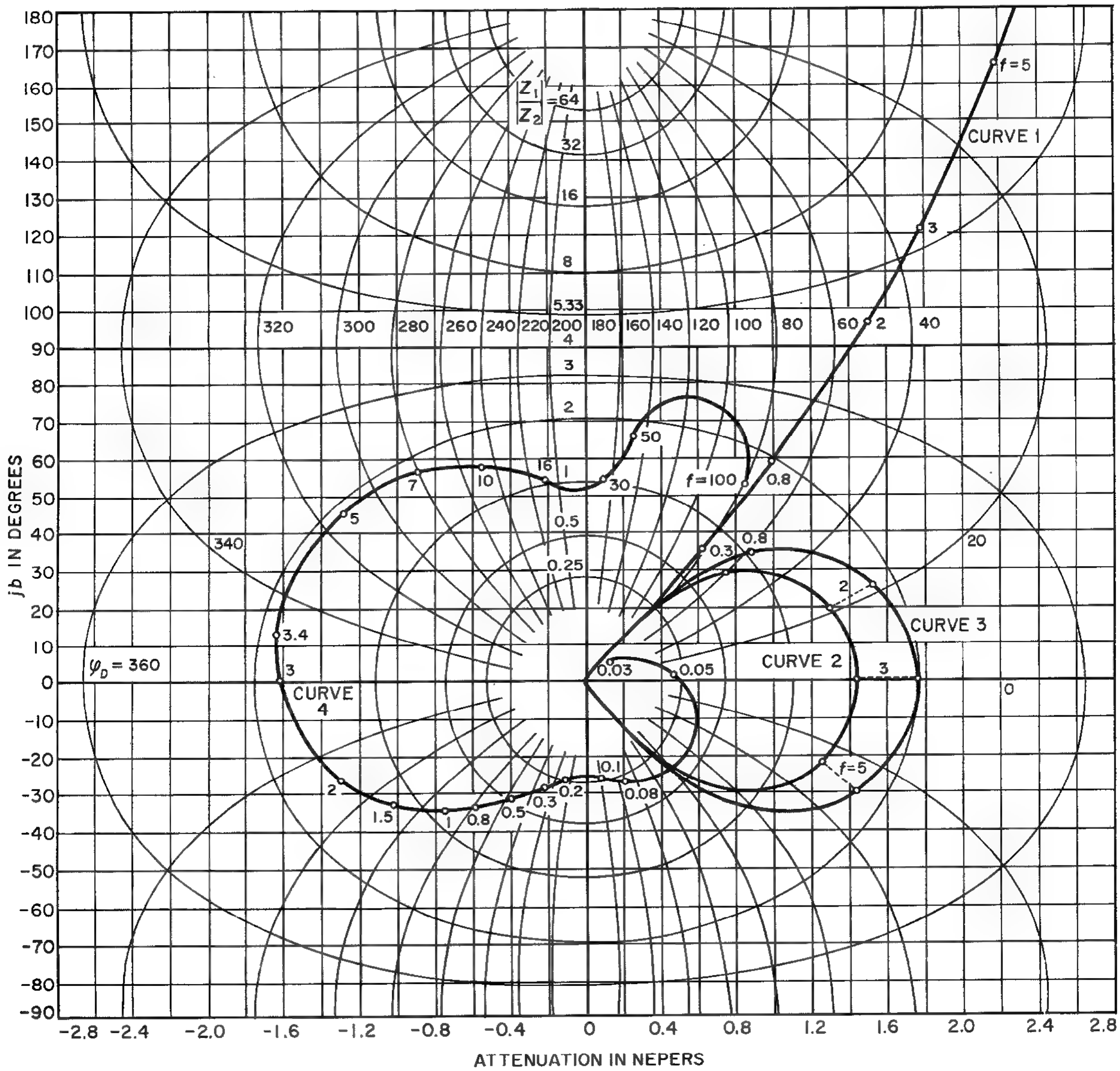
$$g = \ln \frac{1 + \mathbf{N}/2}{1 - \mathbf{N}/2}. \tag{5}$$

Equation (2) is represented in Figure 5 as a diagram of the complex relation of the bridge-network branches $\mathbf{Z}_1/\mathbf{Z}_2 = |\mathbf{Z}_1/\mathbf{Z}_2| \exp j\varphi_D$ with the parameters $\mathbf{Z}_1/\mathbf{Z}_2$ and φ_D in the plane of the image transfer constant $a + jb$ for positive

Figure 5—Diagrams of the functions

$$a + jb = \ln \frac{1 + \frac{1}{2}(\mathbf{Z}_1/\mathbf{Z}_2)^{1/2}}{1 - \frac{1}{2}(\mathbf{Z}_1/\mathbf{Z}_2)^{1/2}}$$

with $\mathbf{Z}_1/\mathbf{Z}_2 = |\mathbf{Z}_1/\mathbf{Z}_2| \exp j\varphi_D$ and for $\mathbf{Z}_1/\mathbf{Z}_2$ and φ_D as parameters. Curve *1* is the propagation constant for a nonloaded cable. Curve *2* is for the circuit of Figure 4 with resistance and capacitance in parallel for $\mathbf{Z}_1$ and in series for $\mathbf{Z}_2$, for frequency limited to 3 kilocycles, and for $\mathbf{Z}_{10}/\mathbf{Z}_{20} = 1.5$. In curve *3*, $\mathbf{Z}_{10}/\mathbf{Z}_{20} = 2$. Curve *4* is for a repeater adjusted for $S_o = 1.6$ nepers at 3 kilocycles. See Figure 25. The circles and figures on the curves designate frequency in kilocycles. The added zero subscript indicates that the impedances are adjusted for maximum power transfer.



root values. It will be seen that, at a given ratio $|\mathbf{Z}_1/\mathbf{Z}_2|$, the maximum attenuation is at $\varphi_D = 0$. The wave attenuation decreases with increasing angular values and reaches zero at $\varphi_D = 180$ degrees. At this point, the image transfer constant is imaginary, as that of an all-pass network. When the difference angles increase to $\varphi_D - 180$ degrees; that is, with negative root values, the wave attenuation becomes negative. In the case under discussion, it is desirable that the network have negative wave attenuation and a positive characteristic impedance. No special dimensioning is necessary to achieve the desired polarity of

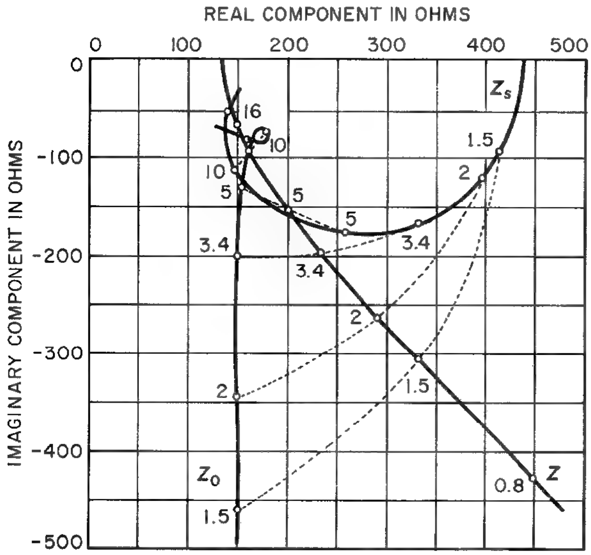


Figure 6—Impedance of 0.03-inch (0.8-millimeter) cable, 3.75 miles (6 kilometers) long, having the characteristics required by the Deutschen Post.

the characteristic impedance. If the termination impedance of an arbitrary symmetrical network equals the positive characteristic impedance of that network, then the network will also have a positive characteristic impedance. If, contrariwise, the termination impedance is equal to the negative characteristic impedance of the network, then the network will also have a negative characteristic impedance. The reason is that reflections are not admissible, whether the network be terminated by positive or by negative characteristic impedances.

The wave attenuation of any quadripole may also be either positive or negative. If the phasors of the three impedances, that is, the bridge-network impedances $\mathbf{Z}_1$ and $\mathbf{Z}_2$ and the characteristic impedance chosen for the termination are located in a semiplane, then the wave attenuation is positive; if the range of a semiplane is exceeded, the wave attenuation becomes negative. The separation line between the two semiplanes can form an arbitrary angle with the real axis. Thus, for instance, a purely resistive line will have negative attenuation if terminated by negative resistors. In the case of our repeater, the termination impedances have positive resistive components; therefore, at least one of the two bridge-network impedances must have a negative resistive component to cause the three impedances to exceed the range of a semiplane.

4. *Replacing Short- and Open-Circuit Impedances by Simple Networks*

For stability reasons, the net loss of the line repeater section must increase quickly outside the transmitted band. The ideal negative line discussed in section 3 does not comply with this requirement. Other branch networks must be examined to obtain one having the desired increase of net loss and, if possible, to require a minimum of circuit components. For instance, the repeater is to be dimensioned for a subscriber cable with a conductor diameter of 0.8 millimeter (0.03 inch) and line constants[2] of

$$R' = 73.2 \text{ ohm per kilometer}$$
$$= 117.8 \text{ ohm per mile}$$
$$C' = 0.038 \text{ microfarad per kilometer}$$
$$= 0.061 \text{ microfarad per mile}$$
$$L' = 0.7 \text{ millihenry per kilometer}$$
$$= 1.13 \text{ millihenry per mile.}$$

The function of this repeater is to compensate for the attenuation of the cable along a section of $2l = 12$ kilometers (7.5 miles) in the band 300 through 3400 cycles (Figure 5, Curve *1*). The open-circuit and short-circuit impedances of half of this section, $l = 6$ kilometers (3.75 miles), are shown in Figure 6. The short-circuit impedance can be approximately balanced in a limited frequency band by a parallel resistance–capacitance branch and the open-circuit impedance by a series resistance–capacitance branch. With

[2] "Telegraphenmessordnung der Deutschen Post," part 4-TMO 4; 1939.

ideal converters, the bridge network of Figure 4 could be constructed with impedance characteristics shown in Figure 7. If R_1 and C_1 are the

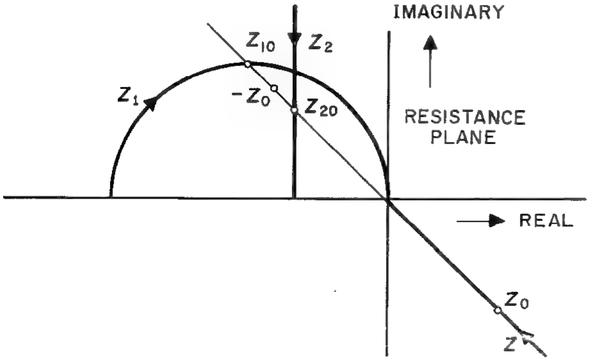


Figure 7—Impedance diagrams of resistance–capacitance branches $\mathbf{Z}_1$ and $\mathbf{Z}_2$, consisting of resistance–capacitance combinations and of the characteristic impedance $\mathbf{Z}$ of the bridge network. $\mathbf{Z}_{10}/\mathbf{Z}_{20} = 1.5$.

components of the branch $\mathbf{Z}_1$, and R_2 with C_2 are those of branch $\mathbf{Z}_2$, the following will be true.

$$\mathbf{Z}_1 = \frac{R_1}{1 + j\omega C_1 R_1} \tag{6}$$

$$\mathbf{Z}_2 = \frac{1 + j\omega C_2 R_2}{j\omega C_2} \tag{7}$$

and the characteristic impedance

$$\mathbf{Z} = \left[\frac{R_1}{j\omega C_2}\frac{1 + j\omega C_2 R_2}{1 + j\omega C_1 R_1}\right]^{\frac{1}{2}}. \tag{8}$$

If the same cutoff frequencies are chosen for both resistance–capacitance components, namely

$$\omega_0 = 1/(C_1 R_1) = 1/(C_2 R_2) \tag{9}$$

the simple relation

$$\mathbf{Z} = (R_1/j\omega C_2)^{\frac{1}{2}} \tag{10}$$

is obtained.

Now the characteristic impedance of the nonloaded cable with the line constants R', L', and C' is

$$\mathbf{Z}_k = [(R' + j\omega L')/(j\omega C')]^{\frac{1}{2}}. \tag{11}$$

Neglecting for the moment the effect of the inductance of the cable, the characteristic impedance of the bridge network and of the cable will coincide if

$$R_1/C_2 = R'/C'. \tag{12}$$

According to (2), the characteristic impedance depends on the ratio of the two bridge network impedances. From (6), (7), and (9), it follows that

$$\frac{\mathbf{Z}_1}{\mathbf{Z}_2} = \frac{j\omega R_1 C_2}{[1 + j(\omega/\omega_0)]^2}. \tag{13}$$

If we write

$$\frac{\mathbf{Z}_1(\omega_0)}{\mathbf{Z}_2(\omega_0)} = \frac{\mathbf{Z}_{10}}{\mathbf{Z}_{20}} = \frac{\omega_0 R_1 C_2}{2} = N_0^2 \tag{14}$$

and

$$\Omega = \omega/\omega_0, \tag{15}$$

then (13) may be rewritten

$$\mathbf{N}^2 = \frac{\mathbf{Z}_1}{\mathbf{Z}_2} = N_0^2\,\frac{2j\Omega}{(1 + j\Omega)^2}. \tag{16}$$

From (16) and (5), the image transfer constant can be computed.

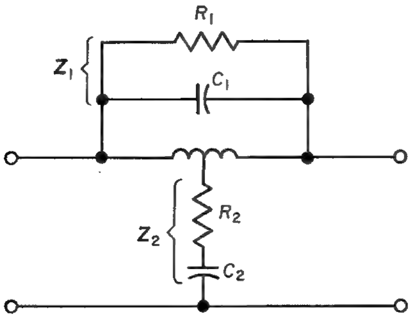
$$a + jb = \ln\frac{1 + \frac{1}{2}N_0\Omega^{\frac{1}{2}} + j(\Omega + \frac{1}{2}N_0\Omega^{\frac{1}{2}})}{1 - \frac{1}{2}N_0\Omega^{\frac{1}{2}} + j(\Omega - \frac{1}{2}N_0\Omega^{\frac{1}{2}})}. \tag{17}$$

The real component of (17), the wave attenuation a, is represented over the scale $\Omega^{\frac{1}{2}}$ of Figure 8 for the values $N_0^2 = \mathbf{Z}_{10}/\mathbf{Z}_{20} = 1.0$, 1.5, and 2.0. The wave attenuation of a nonloaded cable varies in this diagram somewhat like a straight line crossing the zero point. If the frequency $\Omega = 1$ is chosen as the upper cutoff frequency of the transmitted band, the requirements of good equalization and rapid decrease of gain outside this band can be closely approximated.

The characteristics of the image transfer constant with positive real component as per (17) are plotted in Figure 5 (Curves *2* and *3*) for the values $N_0^2 = \mathbf{Z}_{10}/\mathbf{Z}_{20} = 1.5$ and 2. The frequency $f_0 = 3$ kilocycles was chosen as the cutoff frequency of the bridge network impedances. If the bridge network shown in Figure 8 were equipped with ideal converters and the terminations consisted of positive characteristic impedances, the corresponding characteristics of the image transfer constant would have the appearance of curves *2* and *3* of Figure 5, rotated by 180 degrees around the zero point. The combination of the line and repeating network would have a smaller phase constant than the line alone. For the operation of the repeater, it would be essential to reduce the phase constant for the whole transmission band for it is well known that the phase

constant of a nearly lossless line causes at mismatch considerable ripple of the effective attenuation.

The achievable reduction is closely related to the net loss outside the transmitted band, however. The more the compensation for attenuation extends beyond the transmitted or signal band,



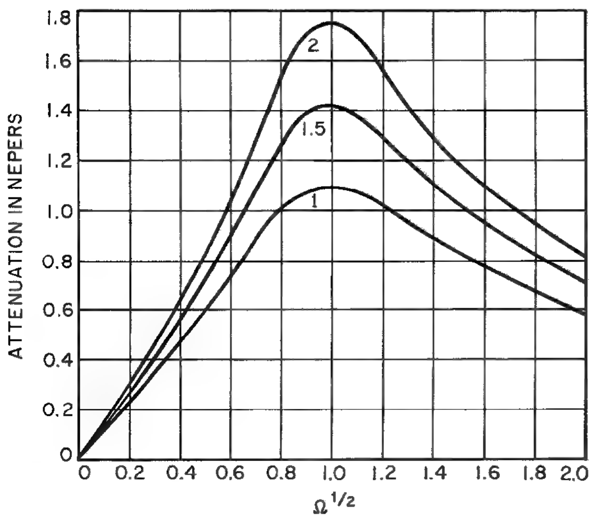


Figure 8—Example of attenuation for a bridge network of resistance and capacitance. $\mathbf{Z} = (R_1/j\omega C_2)^{1/2}$. $C_1R_1 = C_2R_2 = 1/\omega_0$. $\Omega = \omega/\omega_0$. The curves are for the indicated values of $\mathbf{Z}_{10}/\mathbf{Z}_{20}$.

the greater is the phase-rotation effect of the repeater. On the other hand, stability considerations demand that the net loss rise steeply beyond the edges of the signal band. An elimination of the phase constant in the whole frequency band, which would amount to the elimination of the delay time of a pulse, cannot be realized because any practical negative impedance would be transformed into a positive resistance, at least at high frequencies.

The following section contains an estimate of the effect of the phase constant in the particularly unfavorable case of operation where both ends of the repeater section are directly terminated by subscriber stations.

5. *Line Loss of a Repeater Section*

The line loss a_B of a symmetrical quadripole having propagation constant $g = a + jb$, characteristic impedance $\mathbf{Z}$, and terminations $\mathbf{R}_1$ and $\mathbf{R}_2$ can be expressed by several partial losses[3]

$$a_B = a + \ln\left|\frac{\mathbf{R}_1 + \mathbf{Z}}{2(\mathbf{R}_1\cdot\mathbf{Z})^{1/2}}\right| + \ln\left|\frac{\mathbf{R}_2 + \mathbf{Z}}{2(\mathbf{R}_2\cdot\mathbf{Z})^{1/2}}\right| + \ln|1 - \mathbf{r}_1\mathbf{r}_2 \exp|-2g|, \quad (18)$$

where

$$\mathbf{r}_1 = \frac{\mathbf{R}_1 - \mathbf{Z}}{\mathbf{R}_1 + \mathbf{Z}}$$

$$\mathbf{r}_2 = \frac{\mathbf{R}_2 - \mathbf{Z}}{\mathbf{R}_2 + \mathbf{Z}}.$$

The first term a is the wave attenuation, the second is the reflection attenuation at the input, the third is the same attenuation at the output. The fourth term takes account of the effect of the reflected waves and is called the interaction term.

For the above-stated case of operation with two subscriber stations, $\mathbf{R}_1 = \mathbf{R}_2 = \mathbf{R}$ may be written. Equation (18) can thus be rewritten

$$a_B = a + \ln\left|\frac{1 - \mathbf{r}^2 \exp - 2(a + jb)}{1 - \mathbf{r}^2}\right|. \quad (19)$$

Now the subscriber-station impedance has a positive phase angle of about 50 degrees while the characteristic impedance of the cable has a negative angle of about -40 degrees. The differences of the absolute values of both impedances are not very large in midband. If the particularly adverse case of equal impedance values is considered, then the reflection coefficient will be equal to unity as a result of the angle difference of 90 degrees and will have a phase of 90 degrees ($\mathbf{r} = j$). Introducing this value $\mathbf{r} = j$ in (19), the second term will be positive at the values $b = 0, \pi, 2\pi \cdots$ and negative at the values $b = \pi/2, 3\pi/2, 5\pi/2 \cdots$. The limit values of the line loss found with these angular values are plotted in Figure 9 for wave attenuations up to 8.7 decibels (1 neper).

[3] J. Wallot, "Theorie der Schwachstromtechnik," Springer-Verlag, Berlin; 1944.

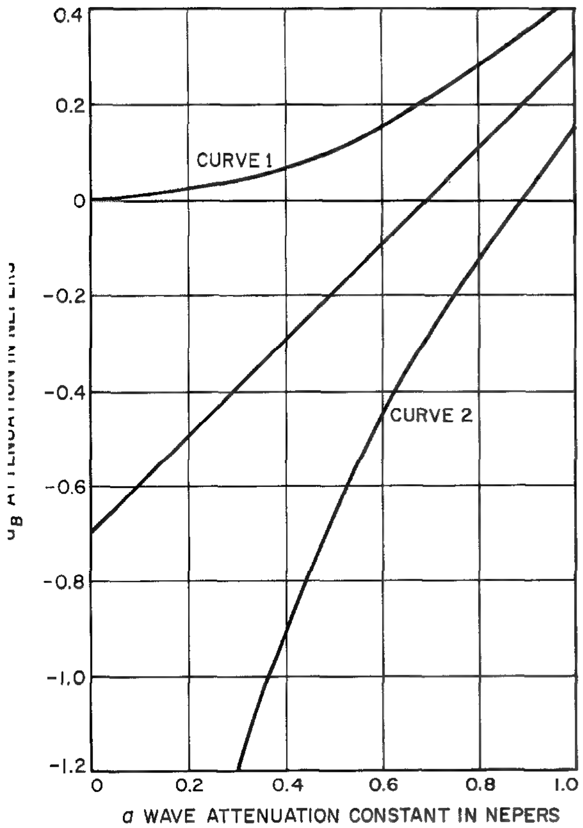


Figure 9—Limiting values of $a_B = a - \ln 2 + \ln|1 - 2\exp - 2(a + jb)|$ the attenuation of a repeater section having end-reflection coefficients corresponding to $r_1 = r_2 = j$. Curve *1* is for the highest values of a_B found at $b = 0, \pi, 2\pi \cdots$. Curve *2* is for the smallest values for $b = \pi/2, 3\pi/2, 5\pi/2 \cdots$. The middle curve is for $a_B = a - \ln 2$.

Figure 9 shows that the line loss can assume large negative values when the wave attenuation a of the repeater section is small. In the discussed case, this gain peak may be expected in the region of 2 kilocycles because at this frequency the phase constant of the repeater section will have the approximate value of $\pi/2$ as can be shown by curves *1* and *2* or curve *3* of Figure 5. The next critical value $3\pi/2$, is far outside the signal band.

6. *Effect of Reflections on Stability*

The varying terminations of the repeater section under operational conditions must not cause self-oscillation. This requirement is fulfilled if the loop gain in the two-wire system is always less than unity, regardless of the phase rotation.[3]

Figure 10 shows the general arrangement of a repeater section. Both ends of the repeater are connected by cables of different lengths to termination impedances. The loop gain depends on the gain $S = \exp s$ of the repeater, on the cable attenuations $A_n = \exp a_n$, and on the reflection coefficients r_n between the characteristic impedances and the termination impedances $\mathbf{Z}_n$. If, for example, the wave U_3 leaves the repeater at point *3*, the junction at this point will reflect the amount U_3r_3 of the wave. The wave entering cable *2* is slightly distorted by the junction point *3*; this will be neglected in the following equation. Now the wave is propagated along the cable several times in the manner indicated in the sketch. The linear addition of the waves returning to point *3* gives

$$\begin{aligned} U'_3 &= U_3[r_3 + (r_4/A_2^2) + (r_4/A_2^2)^2 r_3 \\ &\qquad + (r_4/A_2^2)^3 r_3^2 + \cdots] \\ &= U_3\{r_3 + (r_4/A_2^2)[1 + (r_3r_4/A_2^2) \\ &\qquad + (r_3r_4/A_2^2)^2 + \cdots]\}. \end{aligned} \tag{20}$$

With summation formula

$$1 + x + x^2 + \cdots = 1/(1 - x), \text{ for } |x| < 1, \tag{21}$$

there is obtained

$$\begin{aligned} U_3'/U_3 &= r_3 + (r_4/A_2^2)\{1/[1 - (r_3r_4/A_2^2)]\} \\ &= r_3 + [r_4/(A_2^2 - r_3r_4)]. \end{aligned} \tag{22}$$

The equation for the voltage ratio U_2'/U_2 can be derived from (22) by correspondingly changing

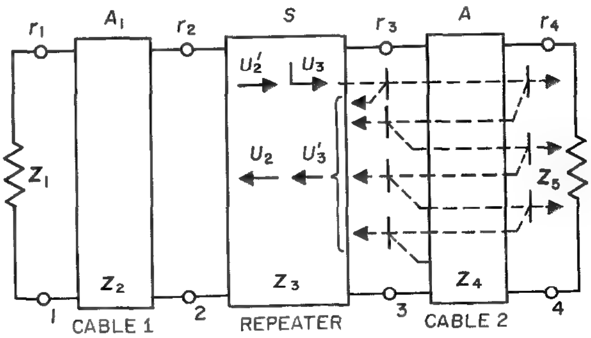


Figure 10—Calculation of loop gain of a repeater section. The cable attenuation $A_n = \exp a_n$, the reflection coefficient $r_n = \left|\dfrac{\mathbf{Z}_n + 1 - \mathbf{Z}_n}{\mathbf{Z}_n + 1 + \mathbf{Z}_n}\right|$ and repeater gain $S = \exp s$.

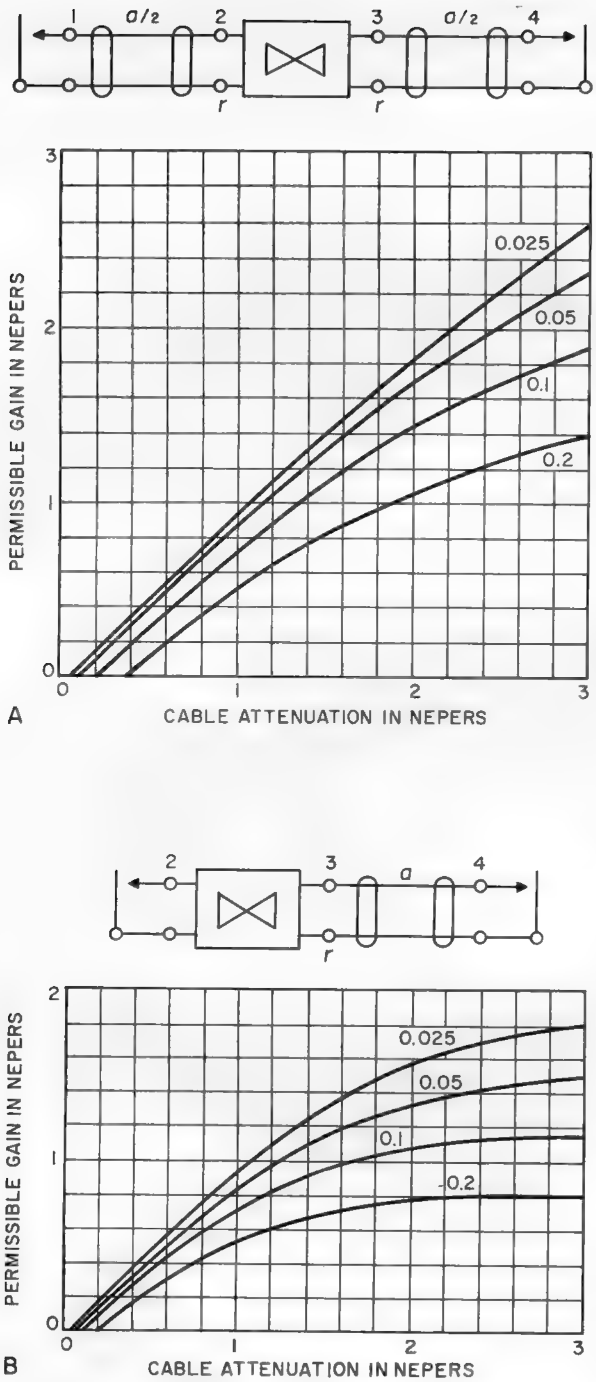


Figure 11—Permissible repeater gain plotted against cable loss for arbitrary open- and short-circuit combinations at the ends of the systems for the four indicated values of reflection coefficient r at the junctions between cable and repeater. The top figure is for the repeater in the middle and the bottom is for the repeater at one end of the cable section.

the suffixes. For the loop gain of 1, appearing at linear addition of the partial voltages, there is obtained the condition

$$S^2(U_2'/U_2)(U_3'/U_3) = 1$$

$$S^2\{r_2 + [r_1/(A_1^2 - r_1 r_2)]\} \times \{r_3 + [r_4/(A_2^2 - r_3 r_4)]\} = 1. \quad (23)$$

Equation (23) was used to establish the characteristics shown in Figure 11. The upper diagram is valid for the repeater installed in the middle of the repeater section, the lower diagram for an end repeater. The reflection coefficient 1 at the ends of the repeater sections was used for both repeaters.

If the reflection factor at one or both ends of the repeater section assumes values greater than 1, then the gain must be reduced at least to the value determined by (23).

The reflection coefficient existing between two arbitrary impedances $\mathbf{Z}_1$ and $\mathbf{Z}_2$ with the ratio $\mathbf{Z}_1/\mathbf{Z}_2 = x \exp j\alpha$ can be computed on the basis of the relation

$$r = (1 - x \exp j\alpha)/(1 + x \exp j\alpha)$$
$$= [(1+x^2-2x\cos\alpha)/(1+x^2+2x\cos\alpha)]^{1/2}. \quad (24)$$

Equation (24) indicates that the reflection coefficient becomes greater than 1 for angles of $\alpha > 90$

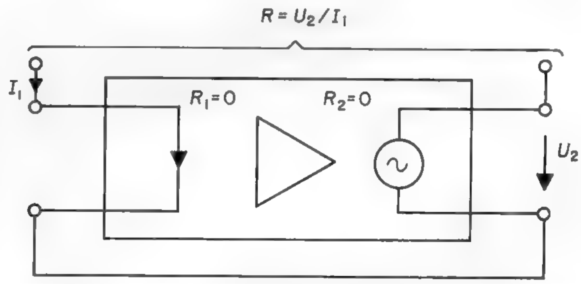


Figure 12—Open-circuit-stable negative impedance. The input current I_1 (origin) produces an output voltage U_2 (result). The input and output resistances of the amplifier are low and connected in series (current feedback).

degrees. If, moreover, both impedances are of the same absolute value ($x = 1$), the reflection factor will assume its maximum value.

In the unfavorable case of a repeater terminated by an inductance, the phase angle of the characteristic impedance of -40 degrees will cause a differential angle of $\alpha = 130$ degrees. When the values are the same ($x = 1$), the

reflection factor will be $r = 2.14$ according to (24).

7. *Properties of Negative Impedances*

Negative impedances can be realized by repeaters with feedback. Positive current feedback

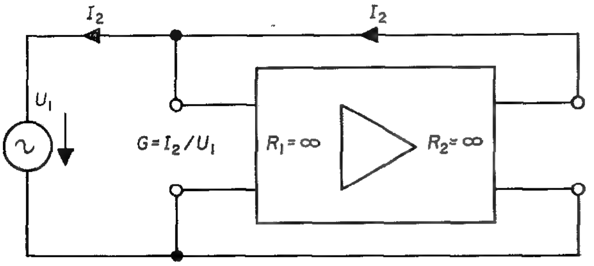


Figure 13—Short-circuit-stable negative impedance. The input voltage U_1 (origin) produces an output current I_2 (result). The amplifier input and output circuits are of high resistance and connected in parallel (voltage feedback).

results in open-circuit-stable impedance and positive voltage feedback in short-circuit-stable impedances. Figures 12 and 13 show simple equivalent circuits of these two types. In the open-circuit-stable type, input and output of the amplifiers are in series (series type), while they are parallel connected in the short-circuit-stable type (parallel type).

If series-type and parallel-type negative impedances are connected in parallel, the loop gain provides negative feedback. This can be shown by Figure 14. Assuming initially the amplification factor to be $\mu_1 = 0$, the right-hand negative impedance is short-circuited through the series-connected input and output of the left-hand

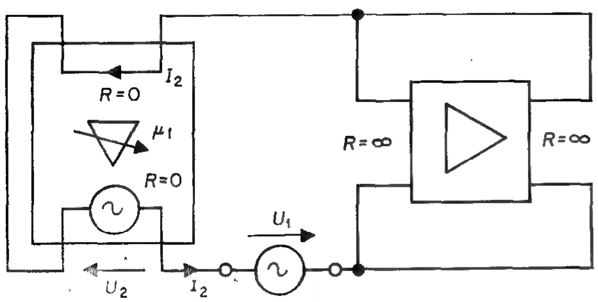


Figure 14—Stability of parallel-connected negative impedances. The test voltage U_1 (origin) produces a current of the direction I_2 (result) which develops a voltage U_2 that counteracts the origin voltage U_1.

negative impedance and thus remains stable. With increasing μ_1, a voltage U_2 is set up that counteracts the origin U_1; in other words, U_2 acts as negative feedback.

This stability of interconnected negative impedances is one of the reasons for connecting a series-type with a parallel-type impedance.

The circle diagram of negative impedances shows that the characteristics proceed partly through the right half of the impedance plane. The basic form of such characteristics will be seen from a simple example. Figure 15 shows a re-

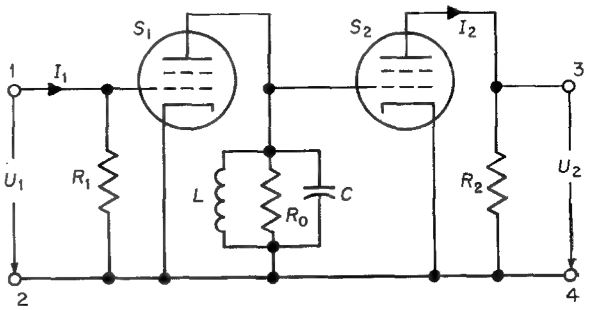


Figure 15—Amplifier producing negative impedance between input and output resistors R_1 and R_2 for cutoff frequencies ω_1 and ω_2. Lower cutoff frequency $\omega_1 = R_0/L$. Upper cutoff frequency $\omega_2 = 1/CR_0$. Transconductance $\mathbf{S} = I_2/\mathbf{Z}_1 = S_0/[1 + j(\omega/\omega_2 - \omega_1/\omega)]$ with $S_0 = S_1S_2R_0$.

peater with the input resistor R_1, the output resistor R_2, and the over-all transconductance

$$\mathbf{S} = I_2/\mathbf{U}_1 = S_0\{1/[1+j(\omega/\omega_2 - \omega_1/\omega)]\} \quad (25)$$

with the cutoff frequencies

$$\omega_1 = R_0/L \quad (26)$$

and

$$\omega_2 = 1/CR_0. \quad (27)$$

If now the repeater input and output are connected in series, the series impedance between terminals *1* and *3* will be

$$\begin{aligned}\mathbf{Z}_1 &= R_1 + R_2 - (I_2R_2/I_1)\\ &= R_1 + R_2 - \mathbf{S}R_1R_2. \quad (28)\end{aligned}$$

When input and output are connected in parallel, the over-all admittance between terminal *2* and the short-circuited terminals *1* and *3* is

$$\frac{1}{\mathbf{Z}_1} = \frac{1}{R_1} + \frac{1}{R_2} - \frac{I_2}{\mathbf{U}_1} = \frac{1}{R_1} + \frac{1}{R_2} - \mathbf{S}. \quad (29)$$

Equations (28) and (29) lead to the equivalent circuits shown in Figures 16 and 17.

The following are valid if the impedances $\mathbf{Z}_1$ and $\mathbf{Z}_2$ are referred to their values at zero frequency.

$$\frac{\mathbf{Z}_1}{R_1 + R_2} = 1 - \mathbf{S}\frac{R_1 R_2}{R_1 + R_2} \tag{30}$$

$$\frac{1/\mathbf{Z}_2}{1/R_1 + 1/R_2} = 1 - \mathbf{S}\frac{R_1 R_2}{R_1 + R_2}. \tag{31}$$

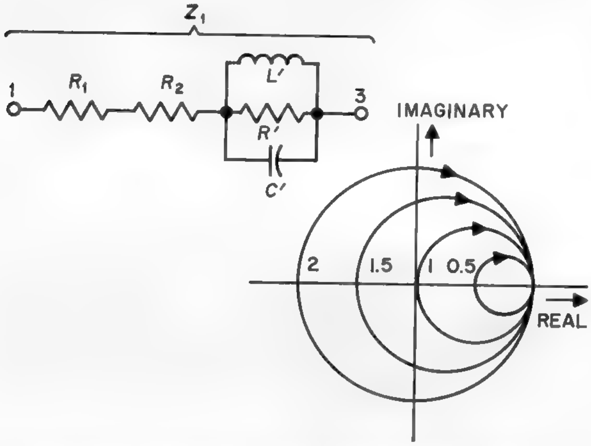


Figure 16—Equivalent circuit and impedance-characteristic diagrams for the indicated values of μ_0 of the amplifier of Figure 15 with input and output in series. $R' = -R_1R_2S_0$. $L' = -L(R_1R_2S_0)/R_0$. $C' = -C(R_0/R_1R_2S_0)$. $\mathbf{Z}_1/(R_1 + R_2) = 1 - \mu_0(\mathbf{S}/S_0)$. $\mu_0 = S_0[R_1R_2/(R_1 + R_2)]$.

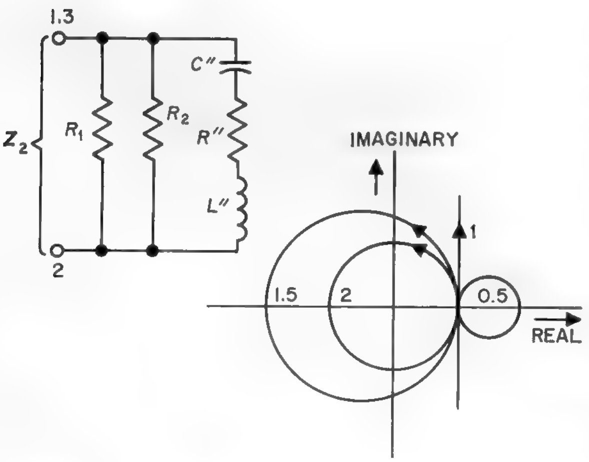


Figure 17—Equivalent circuit and impedance-characteristic diagrams for the indicated values of μ_0 of the amplifier of Figure 15 with input and output in parallel. $R'' = -(1/S_0)$. $L'' = -C(R_0/S_0)$. $C'' = -L(S_0/R_0)$. $\mathbf{Z}_2/[R_1R_2/(R_1 + R_2)] = 1/[1 - \mu_0(\mathbf{S}/S_0)]$. $\mu_0 = S_0[R_1R_2/(R_1 + R_2)]$.

The impedance characteristics of $\mathbf{Z}_1$ and $\mathbf{Z}_2$, again referred to zero frequency, are also shown with the parameter $\mu_0 = S_0(R_1R_2)/(R_1 + R_2)$ in Figures 16 and 17. It will be seen that, with increasing amplification μ_0, the impedance $\mathbf{Z}_1$ passes the zero point and becomes increasingly negative while $\mathbf{Z}_2$ becomes infinitely large before assuming negative values.

8. Construction and Properties of Negative-Impedance Repeaters

Basic diagrams of networks resulting in negative impedances are given in Figures 18 and 19.

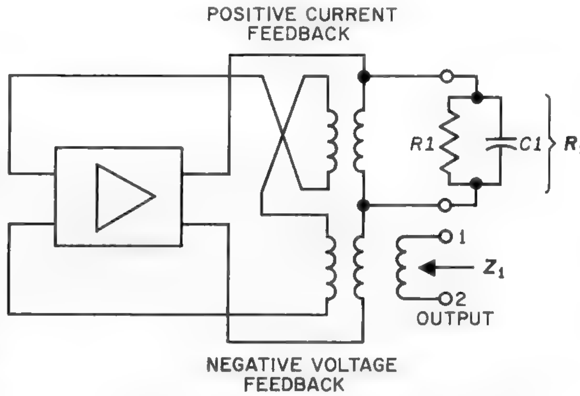


Figure 18—Basic circuit of an open-circuit-stable network in the series branch of the line.

The network destined for the series branch of the bridge network (Figure 18) has positive current feedback; it is of the series type and is, therefore, open-circuit stable; it has been called *NSI* for negative series impedance. The feedback voltage returned from the output renders the negative impedance $\mathbf{Z}_1$ insensitive to gain variations. $\mathbf{Z}_1$ is rotated in its phase by about 180 degrees against the resistance $\mathbf{R}_1$. The value of $\mathbf{Z}_1$ is about proportional to that of $\mathbf{R}_1$. The output of the negative-serves-impedance network will pass direct current.

The network for the shunt branch of the bridge (Figure 19) is of the parallel type and is called *NPI* for negative parallel impedance. This can easily be seen in the case where $\mathbf{R}_2 = 0$. The transformers must be poled so that in the case of $\mathbf{R}_2 = 0$, the input and output of the repeater are coupled by positive feedback. If the center tap of the repeating inductor is connected to the

inductor center and if the amplification is very high, then $\mathbf{Z}_2 = -R_2$. The capacitor in series with the tap renders the negative-parallel-impedance network impassable for direct current.

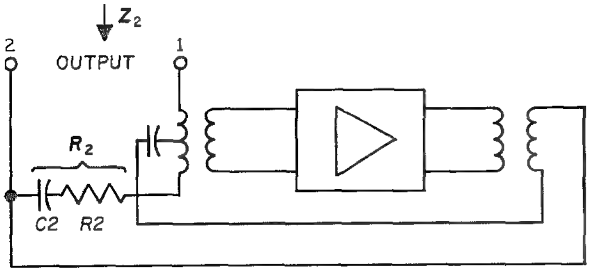


Figure 19—Basic circuit of a short-circuit-stable network in the shunt branch of the line.

transistor and the frequency characteristics of the transformer produce a translation coefficient in the transmitted band that is not constant but is of a complex frequency-dependent magnitude. Thus the negative impedance $\mathbf{Z}_1$ transformed into the line does not have the impedance characteristic of the combination *R1*, *C1* rotated by 180 degrees as in Figure 7, but rather the impedance characteristic in Figure 22. When strong speech and dial signals appear, the rectifier *CR* prevents overloading of the transistor and also reduces the loss of these signals.

The negative-parallel-impedance network producing negative impedance $\mathbf{Z}_2$ is connected by the "*NP* ON" jumper to the center tap of transformer *T1*. Capacitor *C* blocks the network

A simplified circuit diagram of a two-wire repeater with negative impedances is shown in Figure 20. Junction transistors with a power dissipation of 50 milliwatts are used as amplifiers. A 24- or 60-volt battery serves as a direct-current power supply.

The two-wire line is connected to the line windings of transformer *T1* through plug connections. These windings have direct-current resistances of not more than 10 ohms so the loss is small in the case of direct-current transmission. That *T1* winding connected to the base branch produces a negative voltage feedback and the *T4* winding connected in series with the latter produces a current feedback. The finite gain of the

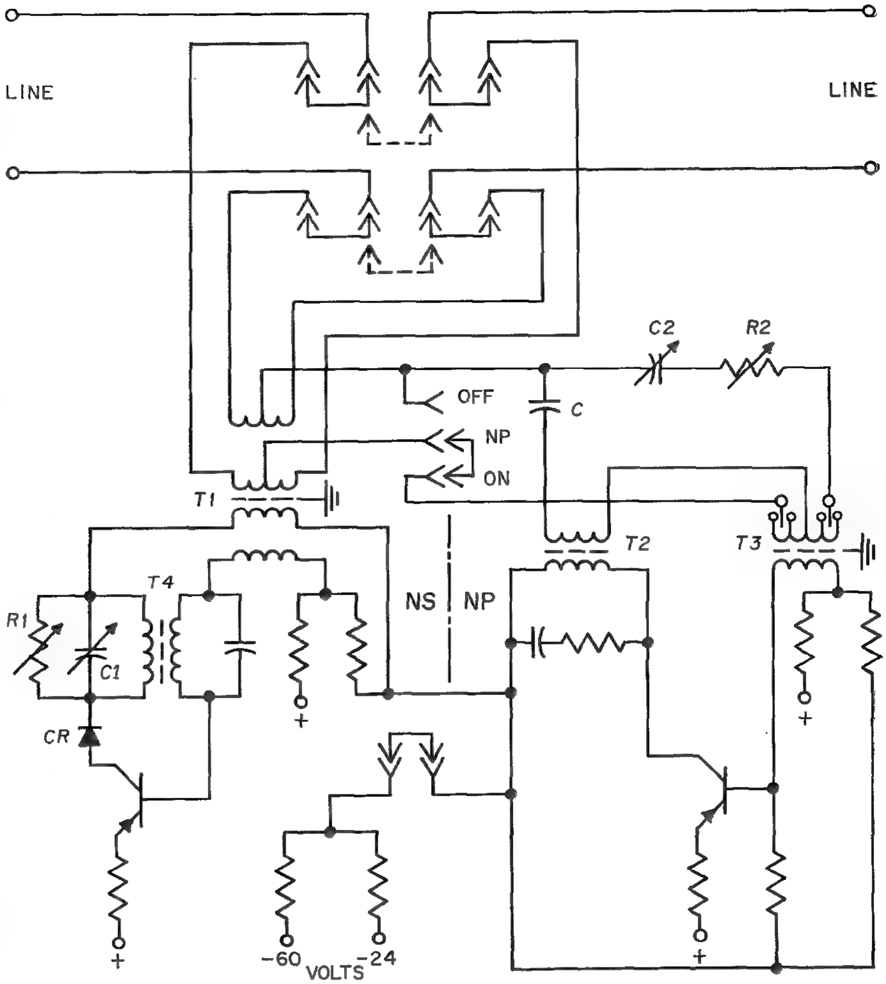


Figure 20—Simplified circuit of the repeater. *NP* and *NS* indicate negative parallel and series impedances, respectively.

against direct current. The characteristic of the negative impedance $\mathbf{Z}_2$ (Figure 22) is also quite different from the $\mathbf{Z}_2$ characteristic in Figure 7; this due to band limitations and to finite loop gain in the negative-parallel-impedance network. The deviations cause both network-loss distortion and reflections. According to Figure 11, the permissible reflection decreases with increasing gain. Therefore it is expedient to match the characteristic impedances of the cable and repeater at that frequency at which the gain is highest. Now, nonloaded cables, because of their inductivity of about $L' = 0.7$ millihenry per kilometer (1.13 millihenries per mile), have a characteristic impedance angle of only 30 to 40 degrees at the highest transmitted frequency.

The adjustment of the negative impedance by the components $R1$, $C1$, $R2$, and $C2$ has to be carried out in a manner differing from the requirement of (9) for the same cutoff frequency inasmuch as the repeater must assume the characteristic impedance of the cable at a frequency near the band limit. The frequency required for this alignment procedure is called the adjustment frequency f_0. The values correlated to this frequency are designated by the same subscript. Then, the requirement of a reflection-free

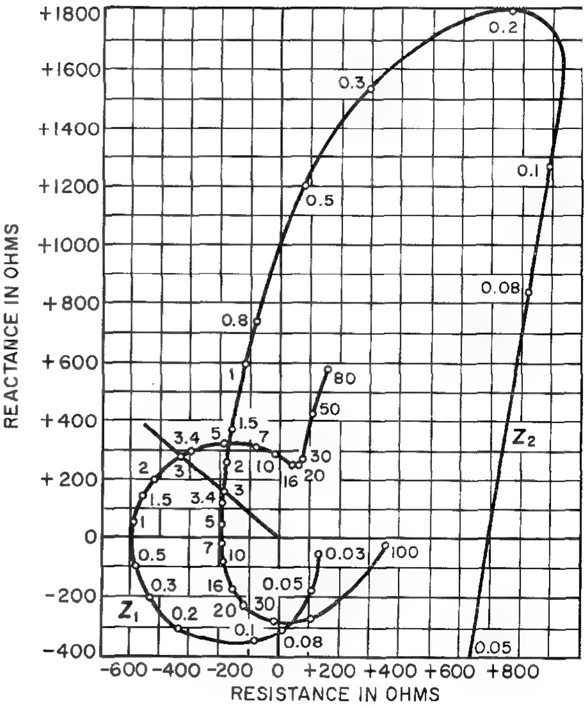
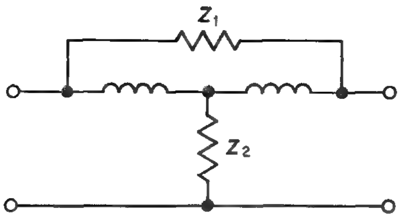


Figure 22—Characteristics of the bridge network impedances $|\mathbf{Z}_1|$ and $|\mathbf{Z}_2|$ with gain adjusted to 1.6 nepers at $f_0 = 3$ kilocycles. The frequency values in kilocycles are indicated by numbers at the various points on the curves.



Z:a

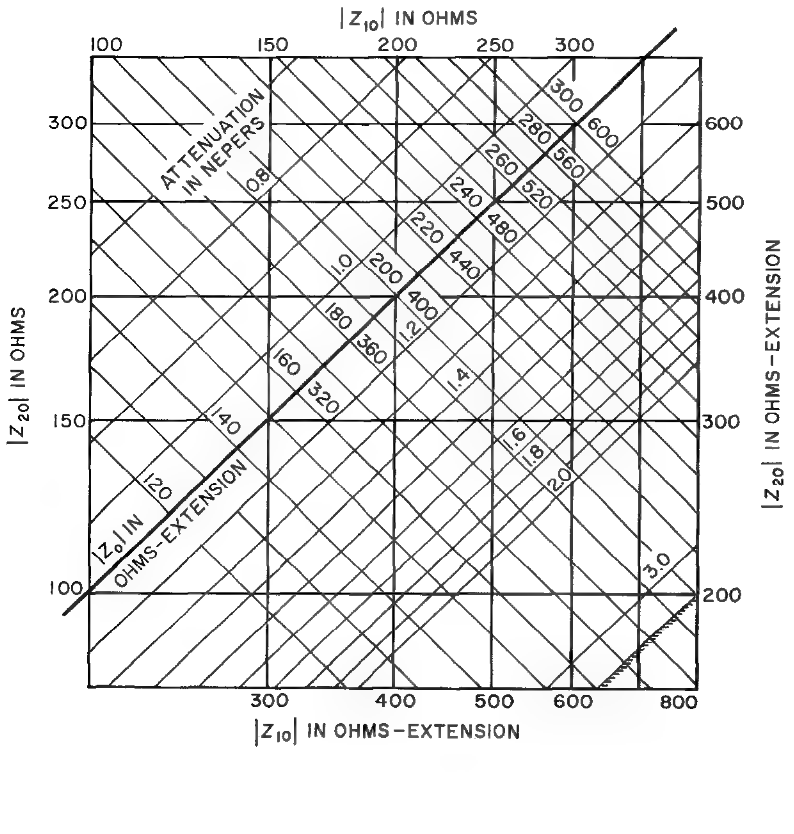


Figure 21 Diagram for determining the bridge network impedances $|\mathbf{Z}_{10}|$ and $|\mathbf{Z}_{20}|$ from the characteristic impedance $|\mathbf{Z}_0|$ and the attenuation a_0 in nepers at frequency f_0. Extension scales for the impedances are given in the lower right half of the chart.

matching of the repeater characteristic impedance Z_0 to the characteristic impedance of the cable, Z_{R0}, at the frequency f_0, may be expressed by the equation

$$\mathbf{Z}_{R0} = \mathbf{Z}_0 = (\mathbf{Z}_{10}\mathbf{Z}_{20})^{1/2}. \quad (32)$$

The factor $\mathbf{N}_0 = (\mathbf{Z}_{10}/\mathbf{Z}_{20})^{1/2}$ determines the gain at frequency f_0. If factor N_0 is chosen as a real one, the adjustment conditions are easily missed;

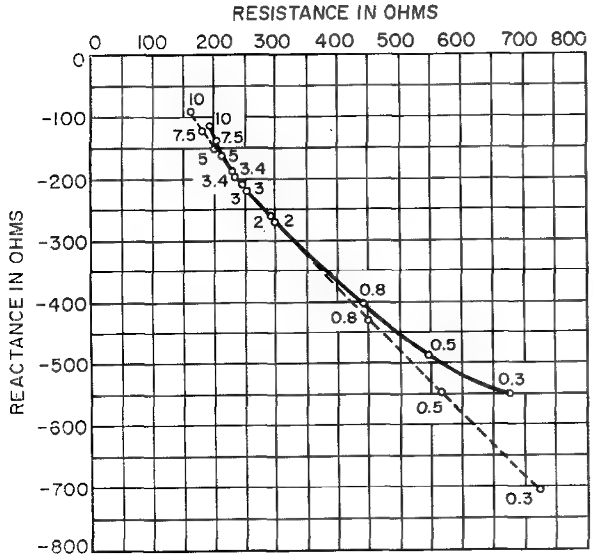


Figure 23—Characteristic impedances of repeater (solid line) and cable (broken line) adjusted to a gain of 1.6 nepers (14 decibels) at 3 kilocycles. The measurement frequencies are indicated along the curves.

however, because the characteristic-impedance angle deviates from 45 degrees, maximum gain is no longer obtained at the adjustment frequency but at a somewhat higher frequency. Therefore it is advantageous to select as the adjustment frequency a value that is smaller than the upper limit of the band, for example, $f_0 = 3.0$ kilocycles.

The factor N_0 from (5) can be found with the aid of the gain required, $S_0 = 1/A_0 = \ln a_0$.

$$N_0 = 2(A_0 - 1)/(A_0 + 1)$$
$$= -2(S_0 - 1)/(S_0 + 1). \quad (33)$$

By means of the characteristic impedance $\mathbf{Z}_0$ required for frequency f_0, we obtain according to (3) and (4) the nominal impedances at frequency f_0.

$$\mathbf{Z}_{10} = N_0\mathbf{Z}_0 \quad (34)$$

and

$$\mathbf{Z}_{20} = \mathbf{Z}_0/N_0. \quad (35)$$

The angles of the negative impedances have to be chosen according to (34) and (35) equal to the angle of the positive characteristic impedance rotated by 180 degrees. Figure 21 shows the nominal values of negative impedances from the amount of characteristic impedance of the cable and from the desired reduction in attenuation.

The two impedance characteristics shown in Figure 22 were measured after the repeater was adjusted at $f_0 = 3.0$ kilocycles, to an attenuation $a_0 = 1.6$ nepers and to the characteristic impedance $Z_0 = 323$ ohms with the angle $\varphi_0 = -40$ degrees.

The characteristic impedance determined by the impedance characteristics of Figure 22 is

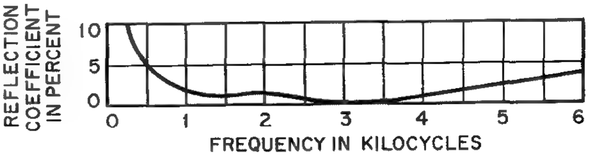


Figure 24—Reflections coefficient of the repeater as derived from Figure 23.

shown in Figure 23. The solid curve is that of the repeater and the broken curve is for the desired characteristic impedance of a cable having an 0.8-millimeter (0.03-inch) core diameter and line constants of

$$R' = 73.2 \text{ ohm per kilometer}$$
$$= 117.8 \text{ ohm per mile}$$
$$C' = 0.038 \text{ microfarad per kilometer}$$
$$= 0.061 \text{ microfarad per mile}$$
$$L' = 0.7 \text{ millihenry per kilometer}$$
$$= 1.13 \text{ millihenry per mile}$$

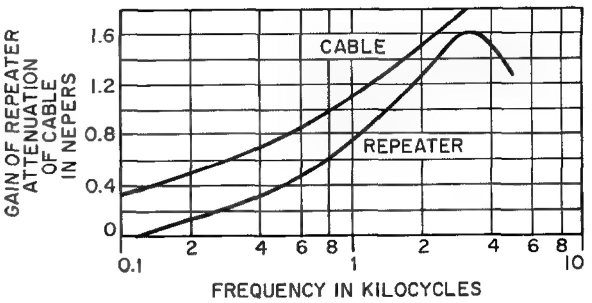


Figure 25—Gain of repeater and attenuation of cable 12 kilometers (7.5 miles) long as a function of frequency. The repeater was adjusted to a gain of 1.6 nepers (14 decibels) at 3 kilocycles.

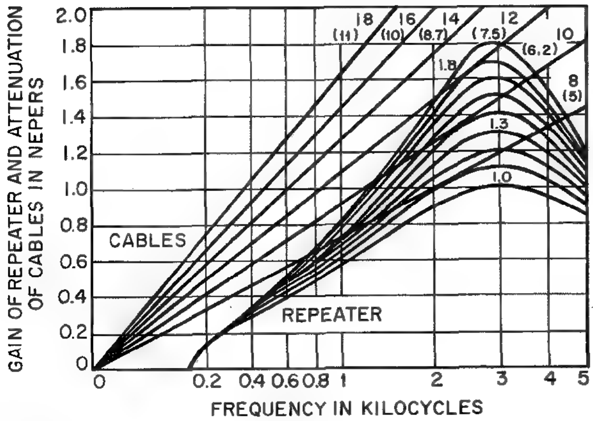


Figure 26—Gain and attenuation characteristics of repeaters and cables. The frequency scale has been dimensioned to make the cable loss linear. The length of each cable is indicated in kilometers (miles). The repeater was adjusted at 2.7 kilocycles to values between 1.0 neper (8.7 decibels) and 1.8 nepers (15.6 decibels) in steps of 0.1 neper.

which is shown for comparison purposes. The reflection coefficient found from both curves is given in Figure 24. The propagation constant was also found from Figure 22 and is shown as curve *4* of Figure 5.

The repeater gain is compared in Figure 25 with the attenuation of a cable 12 kilometers (8 miles) long.

Other gain curves were established using the adjustment frequency f_0 = 2.7 kilocycles (Figure 26). The equalization of the transmitted band is still good at this frequency. If the adjustment frequency chosen is as low as possible, a given cable section can be compensated for at a low value of maximum gain. This lowers the requirements for the reflection attenuation, which are highest at maximum gain.

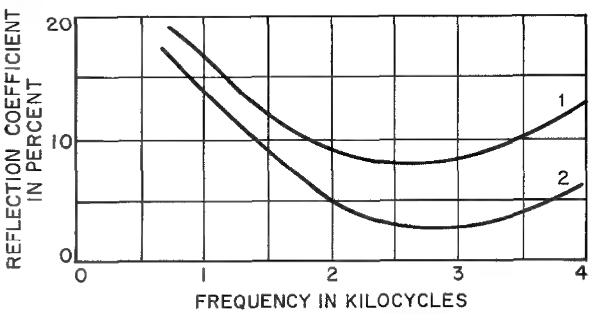


Figure 28—Permissible reflection coefficient of repeaters and cable sections of 14 kilometers (8.7 miles) having a net loss of 0.5 neper (4.3 decibels). Curve *1* is for the repeater in the middle of the line and *2* is for the repeater at the end of the line.

Figure 27 is an example for the net loss of a cable that is 14 kilometers (8.7 miles) long, having an attenuation a_{800} = 1.15 nepers (10 decibels) at 800 cycles. A net loss of 0.5 neper (4.3 decibels) is obtained with the repeater for 800 cycles, the gain being adjusted to 1.55 nepers (13 decibels) at the adjustment frequency

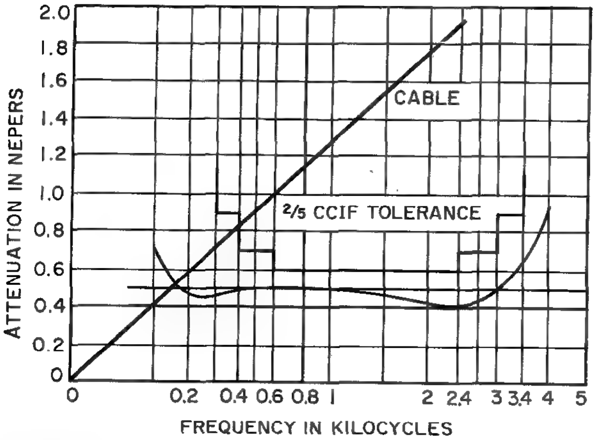


Figure 27—Net loss of repeater section for a cable 14 kilometers (8.7 miles) long. The repeater was adjusted for a gain of 1.55 nepers (13.5 decibels) at 2.7 kilocycles.

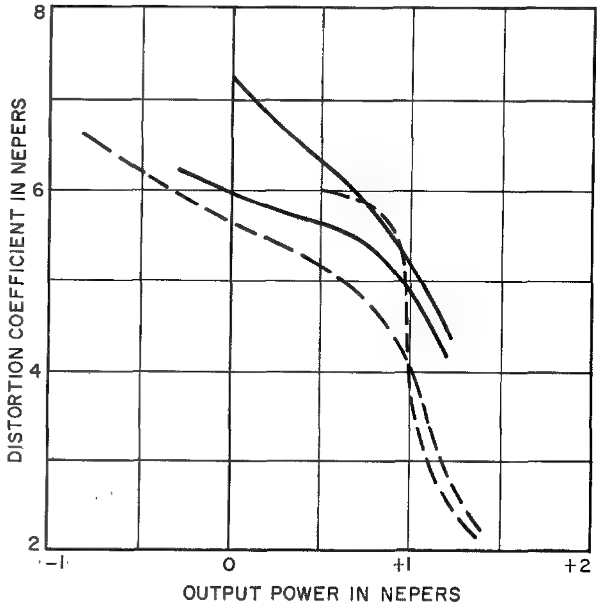


Figure 29—Distortion coefficient of the repeater terminated by cable balancing networks.

$f_0 = 2.7$ kilocycles. The linear distortion, referred to 800 cycles, is within the $\frac{2}{5}$ tolerances recommended by the Comité Consultatif International Téléphonique.

The permissible reflections between the characteristic impedances of the cable and the repeater terminals for this net loss were found in Figure 11 and plotted in Figure 28. These values lead to self-excitation of the repeater section only in those cases where the reflected waves can be added linearly.

Apart from the precise adjustment of its characteristic impedance, the repeater has another feature of importance: its constancy, on which the singing stability of the repeater section depends. When the supply voltage varies by $\pm$ 20 percent, the additional reflection amounts to not more than $\pm$ 0.5 percent, the gain being changed by a maximum of $\pm$ 0.01 neper (0.08 decibel).

The distortion coefficient for a termination by positive characteristic impedances of the cable is shown in Figure 29. The solid curves are valid for a cutoff frequency of 3400 cycles and the broken-line curves for 800 cycles. In the band 300 through 3400 cycles, the point of overload is reached at a voltage level of about + 1.0 neper (8.7 decibels).

Telephone Statistics of the World*

DURING 1956, about 8.8 million telephones were added throughout the world, making a total of 109.8 million on January 1, 1957. More than half of the world's telephones were in the United States, where several thousand privately owned and operated systems provided a telephone for one out of every three individuals. In Europe, which had 29.7 percent of the world's telephones, mostly under public operation, there was a telephone for approximately every 17 individuals.

For the purpose of this compilation, only those telephones that can be connected to a commercial public system are counted. Fourteen countries reported more than 1 million telephones in service January 1, 1957: United States, United Kingdom, Canada, Western Germany, Japan, France, Italy, Sweden, Australia, Switzerland The Netherlands, Argentina, Spain, and Easter[n] Germany. Of the world's principal countries, [9] had more than 15 telephones per 100 of population: United States (35.4), Sweden (31.5) Canada (27.6), New Zealand (25.6), Switzerlan[d] (25.5), Denmark (20.5), Australia (18.5), Norway (17.8), and Iceland (17.4).

TELEPHONES IN CONTINENTAL AREAS

Partly estimated; statistics reported as of other dates have been adjusted to January 1, 1957

Continental Area	Total Telephones			Privately Owned		Automatic (Dial)	
	Number	Percent of Total World	Per 100 Population	Number	Percent of Total	Number	Percent of Total
North America	64 723 700	58.9	34.7	64 046 000	99.0	55 719 800	86.1
Middle America	772 800	0.7	1.3	695 700	90.0	596 500	77.2
South America	2 695 300	2.5	2.1	1 268 600	47.1	2 229 000	82.7
Europe	32 606 600	29.7	5.8	5 218 100	16.0	25 957 000	79.6
Africa	1 546 100	1.4	0.7	22 400	1.4	1 091 500	70.6
Asia	4 929 900	4.5	0.3	3 521 900	71.4	2 889 800	58.6
Oceania	2 525 600	2.3	16.7	178 400	7.1	1 794 700	71.1
World	109 800 000	100.0	4.0	74 951 100	68.3	90 278 300	82.2

New York, with more telephones than any other city, had almost twice as many as Greater London, which ranked second. On a per-capita basis, Washington, District of Columbia, led among the world's large cities with 65.3 telephones per 100 population and Stockholm, Sweden, was first outside the United States with 55.9.

A subdivision in certain of the tables shows the number of telephones operated under private and government ownership. The latter category has reference to municipal and state, as well as national, ownership.

The statistics in this compilation are based on questionnaires sent to the telephone administrations of the various countries throughout the world.

* Abridgement from a booklet issued by the American Telephone and Telegraph Company; New York, New York. After the issuance of the statistics published here, and thus not reflected in them, the government of the Union of Soviet Socialist Republics informed the American Telephone and Telegraph Company that on January 1, 1957, there were 3 366 000 telephones in that country, 1 504 000 of them automatic, and there were 431 000 telephones in Moscow. The corresponding figures for January 1, 1958 were 3 558 000 total, 1 640 000 automatic, and 454 000 in Moscow.

Telephones in Countries of the World as of January 1, 1957

Country	Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
				Private	Government
NORTH AMERICA					
Alaska	30 655	17.52	80.2	7 478	23 177
Canada	4 502 326	27.56	77.3	3 848 137	654 189
Greenland	0	—	—	—	—
St. Pierre and Miquelon	300	6.00	0	0	300
United States	60 190 377	35.45	86.8	60 190 377	0
MIDDLE AMERICA					
Bahamas	7 220	6.20	98.8	0	7 220
Barbados	6 968	3.04	100	6 968	0
Bermuda	8 810	20.02	100	8 810	0
British Honduras	924	1.13	0	45	879
Canal Zone (1) (2)	7 564	28.01	100	0	7 564
Costa Rica	11 245	1.12	2.2	10 856	389
Cuba	143 730	2.37	89.3	143 730	0
Dominican Republic	11 750	0.44	92.9	11 630	120
El Salvador	10 316	0.45	75	0	10 316
Guadeloupe and Dependencies	1 855	0.81	0	0	1 855
Guatemala	11 000	0.32	80	0	11 000
Haiti (3)	4 400	0.13	90	0	4 400
Honduras	5 486	0.33	90	0	5 486
Jamaica and Dependencies	25 000	1.58	97.6	25 000	0
Leeward Islands:					
Antigua	500	0.96	0	0	500
Montserrat	98	0.75	0	0	98
St. Christopher-Nevis	325	0.61	0	0	325
Virgin Islands (United Kingdom)	1	0.01	0	0	1
Total	924	0.72	0	0	924
Martinique	4 143	1.73	70	0	4 143
Mexico	383 257	1.24	73.2	382 609	648
Netherlands Antilles	6 061	3.25	97.1	0	6 061
Nicaragua	5 735	0.45	68.7	0	5 735
Panama	21 635	2.28	82.1	21 080	555
Puerto Rico	65 190	2.84	65.8	60 440	4 750
Trinidad and Tobago	25 431	3.42	87.5	25 431	0
Virgin Islands (United States)	2 780	10.30	0	0	2 780
Windward Islands:					
Dominica	351	0.55	0	0	351
Grenada	1 200	1.50	0	0	1 200
St. Lucia	452	0.51	68.6	0	452
St. Vincent	426	0.53	0	0	426
Total	2 429	0.78	12.8	0	2 429
SOUTH AMERICA					
Argentina	1 155 198	5.87	82.8	82 189	1 073 009
Bolivia	11 700	0.36	91	11 700	0
Brazil	842 800	1.41	82.8	842 800	0
British Guiana	4 819	0.95	14.7	0	4 819
Chile	152 690	2.17	69.8	151 840	850
Colombia	197 752	1.52	94.4	0	197 752
Ecuador	17 000	0.44	60	2 000	15 000
Falkland Islands and Dependencies	391	17.77	0	0	391
French Guiana	779	2.78	0	0	779
Paraguay (3)	6 400	0.40	86	0	6 400
Peru	67 832	0.69	80.3	67 832	0
Surinam	3 883	1.55	94.7	0	3 883
Uruguay (3)	122 600	4.63	75	0	122 600
Venezuela (3)	111 500	1.85	94	110 250	1 250
EUROPE					
Albania (4)	1 555	0.14	10.6	0	1 555
Andorra	100	1.67	0	0	100
Austria	540 524	7.74	87.7	0	540 524
Belgium	931 439	10.41	81.8	0	931 439
Bulgaria (5)	54 347	0.77	39.4	0	54 347
Channel Islands:					
Guernsey and Dependencies	10 660	23.17	27.3	0	10 660
Jersey	14 647	25.70	0	0	14 647
Total	25 307	24.57	11.5	0	25 307
Czechoslovakia (5)	350 708	2.88	59.4	0	350 708
Denmark	922 881	20.52	47.3	813 841	109 040
Finland	486 193	11.27	70.1	372 066	114 127
France	3 313 426	7.57	70.2	0	3 313 426

(1) Excluding telephone systems of the military forces.
(2) June 30, 1956.
(3) Data partly estimated.
(4) January 1, 1943 (latest official statistics).
(5) January 1, 1948 (latest official statistics).
(6) March 31, 1957.
(7) January 1, 1947 (latest official statistics).
(8) Under government operation since 1949.
(9) January 1, 1936 (latest official statistics available during this compilation; but see footnote on page 138).
(10) Includes data for the Isle of Man.

Telephones in Countries of the World as of January 1, 1957—*Continued*

Country	Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
				Private	Government
EUROPE (Continued)					
Germany, Democratic Republic	1 066 582	5.98	92.2	0	1 066 582
Germany, Federal Republic	4 323 225	8.26	96	0	4 323 225
Gibraltar	1 928	7.71	100	0	1 928
Greece	136 835	1.70	93.6	0	136 835
Hungary	347 672	3.55	77.8	0	347 672
Iceland	28 260	17.44	63.6	0	28 260
Ireland	123 619	4.27	69.1	0	123 619
Italy	2 609 127	5.40	95.6	2 609 127	0
Liechtenstein	2 929	19.53	100	0	2 929
Luxemburg	35 540	11.39	84.4	0	35 540
Malta and Gozo (6)	9 170	2.88	0	0	9 170
Monaco	6 953	33.11	100	0	6 953
Netherlands	1 229 174	11.22	96.1	0	1 229 174
Norway (2)	614 523	17.75	66.3	51 862	562 661
Poland	378 000	1.37	68.8	0	378 000
Portugal	279 537	3.15	65 2	188 895	90 642
Rumania (7)	127 153	0.77	75.8	126 131 (8)	1 022
Saar	59 955	6.00	100	0	59 955
San Marino	450	3.21	100	0	450
Spain	1 199 078	4.09	79.2	1 181 437	17 641
Sweden	2 312 223	31.50	78.7	0	2 312 223
Switzerland	1 293 743	25.50	99.6	0	1 293 743
Turkey	173 730	0.70	86.8	0	173 730
Union Soviet Socialist Republics (9)	861 181	0.52	19.9	0	861 181
United Kingdom (6) (10)	7 218 791	14.04	78.2	0	7 218 791
Yugoslavia	175 341	0.98	70.1	0	175 341
AFRICA					
Algeria	144 402	1.41	79.6	0	144 402
Angola	3 802	0.09	95.9	0	3 802
Ascension Island	44	22.00	68.2	44	0
Basutoland	770	0.12	5	0	770
Bechuanaland	254	0.08	0	0	254
Belgian Congo	18 046	0.14	78	0	18 046
Cameroons (French Administration)	3 616	0.11	58.1	0	3 616
Cape Verde Islands	127	0.07	0	0	127
Comoro Islands	0	—	—	—	—
Egypt	179 988	0.75	78.8	0	179 988
Ethiopia and Eritrea	7 264	0.04	80.5	0	7 264
French Equatorial Africa	5 735	0.12	36.5	0	5 735
French West Africa	25 351	0.13	57.1	0	25 351
Gambia	489	0.17	99.2	0	489
Ghana (6)	15 821	0.34	40.3	0	15 821
Ifni	121	0.27	0	121	0
Kenya	29 447	0.48	76.5	0	29 447
Liberia (3)	1 500	0.12	100	425	1 075
Libya (3)	7 500	0.67	77	0	7 500
Madagascar and Dependencies	10 284	0.21	46.1	1 279	9 005
Mauritius and Dependencies	7 282	1.26	8	0	7 282
Morocco	124 893	1.26	83.2	19 533	105 360
Mozambique	8 271	0.14	75.1	0	8 271
Nigeria, Federation of, and British Cameroons	24 536	0.07	48.4	0	24 536
Portuguese Guinea	346	0.06	0	0	346
Reunion	4 649	1.67	0	0	4 649
Rhodesia and Nyasaland:					
Northern Rhodesia	14 113	0.64	93.8	1 477	12 636
Nyasaland	3 604	0.14	89.6	0	3 604
Southern Rhodesia	54 579	2.16	81.7	0	54 579
Total	72 296	0.98	84.4	1 477	70 819
Ruanda-Urundi	1 071	0.03	92.3	0	1 071
St. Helena	112	2.24	0	0	112
São Tomé and Principe	312	0.52	0	0	312
Seychelles and Dependencies	160	0.41	100	160	0
Sierra Leone	2 257	0.09	81.1	0	2 257
Somaliland, British Protectorate	330	0.05	0	0	330
Somaliland, French	762	1.17	100	0	762
Somaliland (Italian Administration)	1 228	0.10	0	0	1 228
South West Africa	10 614	2.23	42.2	0	10 614
Spanish Guinea	827	0.40	71.2	827	0
Spanish North Africa	6 199	4.30	100	6 199	0
Spanish Sahara	40	0.08	0	40	0
Sudan	18 569	0.18	80.2	0	18 569
Swaziland	997	0.45	35.5	0	997
Tanganyika	11 462	0.14	53.8	0	11 462

Country		Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership: Private	Ownership: Government
AFRICA (continued)						
Togoland		1 095	0.09	68.1	0	1 095
Tunisia		33 710	0.88	57.3	0	33 710
Uganda		10 697	0.19	75.8	0	10 697
Union of South Africa	(6)	765 540	5.40	67.4	0	765 540
Zanzibar and Pemba		1 045	0.38	4.3	0	1 045
ASIA						
Aden Colony		3 165	2.26	100	0	3 165
Aden Protectorate		0	—	—	—	—
Afghanistan	(3)	6 200	0.05	30	0	6 200
Bahrain		1 756	1.40	100	1 756	0
Bhutan		0	—	—	—	—
Brunei		160	0.29	0	0	160
Burma	(3)	7 400	0.04	0	0	7 400
Cambodia		2 637	0.06	0	0	2 637
Ceylon		28 757	0.32	93.8	0	28 757
China	(5)	244 028	0.05	72.9	94 945 (7)	149 083
Cyprus		12 913	2.43	83.8	0	12 913
Hong Kong		63 760	2.56	100	63 760	0
India	(6)	314 885	0.08	52.6	3 528	311 357
Indonesia		79 227	0.09	12.1	0	79 227
Iran		63 927	0.29	55	0	63 927
Iraq	(6)	41 725	0.83	77.2	0	41 725
Israel		72 445	3.87	92.1	0	72 445
Japan	(6)	3 486 821	3.84	55.7	3 486 821	0
Jordan		11 034	0.74	72.4	0	11 034
Korea, Republic of		50 256	0.23	40.3	0	50 256
Kuwait		1 800	0.87	80	0	1 800
Laos		536	0.03	51.1	0	536
Lebanon		38 497	2.65	90.5	0	38 497
Macao		1 865	0.93	100	0	1 865
Malaya		57 358	0.90	66	0	57 358
Maldive Islands		0	—	—	—	—
Muscat and Oman		152	0.03	100	152	0
Nepál		0	—	—	—	—
Netherlands New Guinea	(3)	1 100	0.16	0	0	1 100
North Borneo		1 705	0.44	81.8	0	1 705
Pakistan		49 892	0.06	68.8	0	49 892
Philippine Republic		63 400	0.28	68.7	56 400	7 000
Portuguese India		266	0.04	0	0	266
Portuguese Timor		443	0.09	0	0	443
Qatar		443	1.11	100	443	0
Ryukyu Islands	(1)	3 924	0.48	37	0	3 924
Sarawak		1 931	0.31	63.4	0	1 931
Saudi Arabia		13 835	0.20	4.3	0	13 835
Singapore		41 672	3.23	100	0	41 672
Syria		36 650	0.91	83.4	0	36 650
Taiwan		47 112	0.50	49.4	0	47 112
Thailand		11 832	0.05	100	0	11 832
Trucial Oman		0	—	—	—	—
Viet-Nam, Republic of		12 054	0.09	85.1	0	12 054
Yemen		0	—	—	—	—
OCEANIA						
Australia		1 762 173	18.48	71	0	1 762 173
British Solomon Islands		208	0.20	0	0	208
Caroline Islands		191	0.44	0	0	191
Cocos (Keeling) Islands		59	9.08	100	0	59
Cook Islands		174	1.09	0	82	92
Fiji Islands		4 327	1.25	58.9	0	4 327
French Oceania		888	1.22	0	0	888
Gilbert and Ellice Islands		108	0.26	74.1	77	31
Guam		10 233	14.83	93.6	0	10 233
Hawaii		178 165	30.10	99.8	178 165	0
Mariana Islands (less Guam)		350	5.00	71.4	0	350
Marshall Islands		485	3.46	99	0	485
Nauru		0	—	—	—	—
New Caledonia and Dependencies		2 504	3.85	64.7	0	2 504
New Hebrides Condominium		250	0.46	0	0	250
New Zealand	(6)	568 339	25.59	62.3	0	568 339
Niue Island		63	1.26	0	0	63
Norfolk Island		50	5.00	0	0	50
Papua and New Guinea		3 883	0.22	51.4	120	3 763
Pitcairn Island		0	—	—	—	—
Samoa, American		331	1.66	100	0	331
Samoa, Western		690	0.71	0	0	690
Tokelau Islands		0	—	—	—	—
Tonga (Friendly) Islands		535	0.94	0	0	535

TELEPHONE CONVERSATIONS FOR THE YEAR 1955

Conversation data were not available for all countries

Country		Number of Conversations in Thousands			Conversations Per Capita
		Local	Toll	Total	
Alaska		106 300	800	107 100	630.0
Algeria		72 300	26 200*	98 500	9.9
Argentina		3 532 000	41 200	3 573 200	183.5
Australia		1 285 400	103 100	1 388 500	147.3
Belgium		522 400	91 100	613 500	68.7
Brazil		3 926 600	52 200	3 978 800	66.5
Canada		7 559 500	171 300	7 730 800	480.7
Ceylon		60 800	4 900	65 700	7.5
Channel Islands		17 600	200	17 800	172.8
Chile		378 500	23 100	401 600	57.9
Colombia		690 400	9 700	700 100	54.1
Cuba		502 000	6 400	508 400	83.7
Denmark		1 056 400	183 300	1 239 700	277.6
Egypt		450 000	12 700	462 700	19.8
El Salvador		21 200	2 400	23 600	10.4
Finland		593 800	83 000	676 800	157.8
France		2 053 200	571 500	2 624 700	60.2
French West Africa		15 600	1 600	17 200	0.9
Germany, Democratic Republic		772 900	119 300	892 200	50.3
Germany, Federal Republic		2 733 900	636 800	3 370 700	66.6
Greece		329 400	6 800	336 200	41.9
Hawaii		294 900	2 400	297 300	530.9
Hungary		438 300	25 900	464 200	47.4
Iceland		63 200	1800	65 000	403.7
Ireland		91 800	14 600	106 400	36.8
Israel		122 500	5 200	127 700	70.4
Italy		4 135 000	268 200*	4 403 200	91.3
Jamaica		55 600	900	56 500	36.1
Japan	(1)	8 520 000	712 900	9 232 900	102.6
Lebanon		51 800	4 600	56 400	38.9
Madagascar		10 700	1 100	11 800	2.4
Malaya		140 800	17 000	157 800	25.2
Mexico		820 500	12 000	832 500	27.3
Morocco		95 700	16 500*	112 200	11.4
Netherlands		872 800	302 600	1 175 400	108.0
Norway	(2)	484 500	58 000	542 500	157.5
Peru		245 800	3 900	249 700	25.9
Philippines		425 500	1 100	426 600	19.2
Portugal		257 000	48 600	305 600	34.6
Puerto Rico		130 400	2 800	133 200	58.5
Saar		80 900	1 400	82 300	82.3
Singapore		148 800	800	149 600	118.4
South West Africa		10 900	1 600	12 500	26.7
Spain		2 770 000	93 700	2 863 700	98.1
Sweden	(3)	3 222 100	120 600	3 342 700	455.3
Switzerland		516 700	444 500*	961 200	191.4
Syria		94 700	6 100	100 800	25.4
Trinidad and Tobago		78 400	4 600	83 000	113.4
Tunisia		26 300	6 900	33 200	8.8
Turkey		228 000	9 200	237 200	9.6
Union of South Africa	(1)	873 500	59 800	933 300	67.1
United Kingdom	(1)	3 781 000	323 000	4 104 000	80.1
United States		68 545 000	3 080 000	71 625 000	425.7
Viet-Nam, Republic of		12 200	200	12 400	0.9
Yugoslavia		285 000	22 600	307 600	17.3

(1) Year ended March 31, 1957.
(2) Year ended June 30, 1956.
(3) Year ended June 30, 1957
* Three-minute units

United States Patents Issued to International Telephone and Telegraph System; November 1957–January 1958

Between November 1, 1957 and January 31, 1958, the United States Patent Office issued 50 patents to the International System. The names of the inventors, company affiliations, subject, and patent numbers are listed below.

H. H. Adelaar, Bell Telephone Manufacturing Company (Antwerp), Binary Electrical Counting Circuit, 2 812 134.

E. Albert, Bell Telephone Manufacturing Company (Antwerp), Measuring Instrument, 2 817 817.

R. A. Andre and L. J. G. Nys, Bell Telephone Manufacturing Company (Antwerp), Matrix for Detachably Mounting Electrical Components, 2 821 691.

M. Arditi, G. A. Deschamps, and J. Elefant, Federal Telecommunication Laboratories, Microwave Filters, 2 819 452.

M. Arditi, G. A. Deschamps, and J. Elefant, Federal Telecommunication Laboratories, Microwave Filters, 2 820 206.

H. Bartels, Süddeutsche Apparatefabrik (Nürnberg), Method of Making Dry-Contact Rectifiers, Particularly Selenium Rectifiers, 2 819 436.

H. Bartels and H. Fritsch, Süddeutsche Apparatefabrik (Nürnberg), Process for Connecting a Tantalum Electrode Pin to an Electrode Body, 2 819 961.

A. H. W. Beck and A. B. Cutting, Standard Telephones and Cables Limited (London), Ionization Manometers, 2 817 030.

G. R. Clark, Federal Telecommunication Laboratories, Selective Gate System for Radar Beacons, 2 815 504.

E. C. L. deFaymoreau, Federal Telecommunication Laboratories, Electromechanical Delay Device, 2 815 490.

E. C. L. deFaymoreau, Federal Telecommunication Laboratories, Signaling System, 2 815 507.

M. Dishal and M. Rogoff, Federal Telecommunication Laboratories, Continuous-Wave Beacon System, 2 817 082.

C. L. Estes, Federal Telecommunication Laboratories, Electrical Signal-Translating System, 2 815 486.

L. Goldstein, M. A. Lampert, and J. F. Heney, Federal Telecommunication Laboratories, Electromagnetic Wave Generator, 2 817 045.

H. Grottrup, C. Lorenz A. G. (Stuttgart), Sensing Arrangement for Stored Information Concerning Positioning of a Mechanical Element, 2 820 216.

S. J. Harris and J. A. Henderson, Capehart–Farnsworth Company, Television Film Scanners Having Sprocket Hole Sensing Means Responsive to Film Refraction Between Holes, 2 818 467.

W. Hauer, Mix & Genest (Stuttgart), Dispatch-Tube System, 2 815 183.

H. L. Horwitz and G. L. Hasser, Federal Telephone and Radio Company, Two-Party-Line Individual Metering, 2 820 848.

J. F. Houdek, Jr., Kellogg Switchboard and Supply Company, Selectively Amplifying Loud-Speaking Telephone, 2 821 572.

T. M. Jackson and E. A. F. Sell, Standard Telephones and Cables Limited (London), Stepping Circuit Arrangement Using Trigger Devices, 2 814 762.

R. V. Judy, Kellogg Switchboard and Supply Company, Dual-Switch Finder Combination, 2 821 573.

K. A. Karow, Kellogg Switchboard and Supply Company, Digit-Absorbing Selector, 2 818 469.

M. Kenmoku, Nippon Electric Company, Limited (Tokyo), Microwave Discharge Tube, 2 814 756.

W. Klein and W. Fritz, C. Lorenz A. G. (Stuttgart), Traveling-Wave-Tube Arrangement, 2 812 469.

K. Klinkhammer, K. Steinbuch, G. Merk, and H. Bretschneider, Mix & Genest (Stuttgart), Digital Register for Communication System, 2 820 849.

G. Kratt, C. Lorenz A. G. (Stuttgart), Translating Device for Type Printing Machines, 2 821 570.

G. Kratt and O. Holstein, C. Lorenz A. G. (Stuttgart), Telegraph Signal Translation Mechanism, 2 820 094.

C. C. Larson, Farnsworth Electronics Company, Jump Compensation for Continuous Motion Film Projection, 2 818 466.

P. E. Lighty, Federal Telecommunication Laboratories, Selenium Rectifier, 2 815 475.

F. Malsch, C. Lorenz A. G. (Stuttgart), Tuning-Indicator Valve of Small Dimensions and High Sensitivity, 2 820 916.

E. M. S. McWhirter and F. W. Warden, Intelex Systems Incorporated, Memory System, 2 814 440.

M. Mattheyses, Federal Telephone and Radio Company, Rectifier Stack, 2 819 434.

R. C. Miles, Federal Telephone and Radio Company, Moving-Target Indicator, 2 818 561.

F. H. Mittag and H. Ringhandt, Standard Elektrik A. G. (Stuttgart), Dispatch Tube System, 2 815 182.

H. Nopp and K. Wernick, Mix & Genest (Stuttgart), Circuit Arrangement for Loudspeaker Telephone Systems, 2 820 096.

R. L. Plouffe, Jr., Federal Telecommunication Laboratories, Delay-Line Pulse Shaper, 2 820 909.

M. C. Poylo, Federal Telecommunication Laboratories, Information Transmission System, 2 815 400.

E. A. Richards and L. J. Ellison, Standard Telephones and Cables Limited (London), Dry Contact Rectifiers, 2 817 047.

E. Richert, Mix & Genest (Stuttgart), Arrangement for the Stopping and Dispatching of Carrier Capsules in Pneumatic Tube Systems, 2 816 719.

T. P. Robinson and B. W. Glover, Standard Telephones and Cables Limited (London), Electrical Control Circuits, 2 821 679.

R. D. Salmon and F. J. L. Turner, Creed & Company, Limited (Croydon), Telegraph-Code Perforating Apparatus, 2 818 116.

K. Sass and K. Klinkhammer, Mix & Genest, Circuit Arrangement for Final Selectors with Three Line Groups Arranged in One Plane, 2 821 574.

W. Sichak, Federal Telecommunication Laboratories, Dual Polarization Antenna, 2 820 965.

E. Stein, Federal Telecommunication Laboratories, Waveform Monitor, 2 817 043.

D. L. Thomas, Standard Telephones and Cables Limited (London), Two-Way Repeaters, 2 812 388.

H. E. Thomas, Federal Telecommunication Laboratories, Sweep Wave Generators, 2 819 392.

L. R. Ullery and R. K. Orthuber, Capehart–Farnsworth Company, Radiation Detector, 2 818 511.

C. Weill, C. Hannigsberg, and H. Adelaar, Bell Telephone Manufacturing Company (Antwerp), Control Circuit for Pulse Generator, 2 820 141.

H. Wolfson, Standard Telephones and Cables Limited (London), Germanium Diodes, 2 818 537.

W. W. Wright, P. Welch, and D. A. Day, Standard Telephones and Cables Limited (London), Electron Discharge Device, 2 815 464.

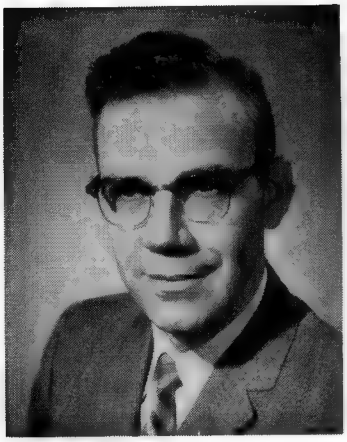
Memory System

2 814 440

E. M. S. McWhirter and F. W. Warden

A data-card storage and reading system is described. The storage cabinet has a record transfer mechanism for feeding the cards from a particular drawer or tray past heads reading the data on the cards, which are then replaced in their drawer.

Contributors in This Issue



Dean W. Davis

Dean W. Davis was born in 1921 in Fort Wayne, Indiana. He received a degree of Bachelor of Science in electrical engineering from Purdue University in 1943. He then served for three years in the United States Army Signal Corps.

He joined the research department of the Farnsworth Electronics Company in 1946, working on special-purpose tubes. He has been in charge of storage-tube development since its inception at Farnsworth in 1949 and now heads the special-tube development laboratory. He reports in this issue on the Iatron storage tube.

Mr. Davis is a member of the Institute of Radio Engineers.

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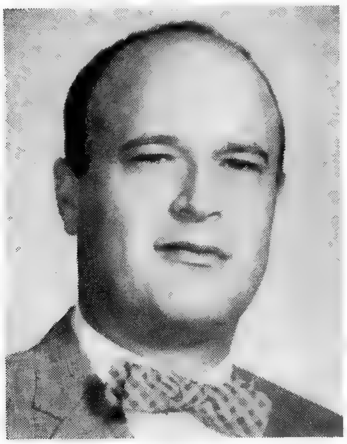

Harry W. Gates

Harry W. Gates was born on May 18, 1921 in Tulsa, Oklahoma. He received BS and MS degrees in electrical engineering from the University of New Mexico in 1949.

From 1942 to 1946, he served as a radar and communications officer in the United States Marine Corps. From 1949 to 1952, he taught ultra-high-frequency theory and did research at the University of New Mexico and the University of Florida.

In 1952, Mr. Gates joined Farnsworth Electronics Company, where he is a senior engineer concerned with air-traffic-control systems design. His paper in this issue describes the Farnsworth Iatron storage tube as used in an indicator for air-traffic control.

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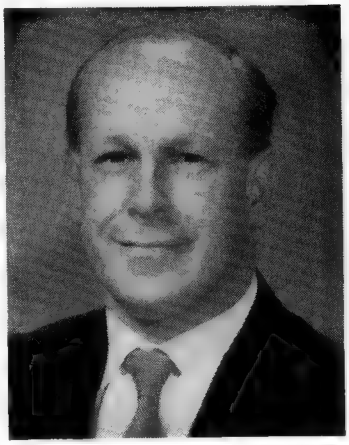

D. S. Girling

D. S. Girling was educated at Leicester City Boys' Grammar School. As a member of the Territorial Army he was called up at the outbreak of war and served with the British Expeditionary Force in their campaign in Belgium in May 1940. He was later commissioned in Royal Signals and was responsible for the training of line mechanics on line carrier equipment.

In May 1946, he joined the capacitor department of Standard Telephones and Cables at North Woolwich. He took charge of the development laboratory in 1948, since when his main interests have been the statistics of capacitor failure, on which he reports in this issue, and techniques for the manufacture of miniature capacitors. He has recently taken charge of the capacitor engineering department. He is an Associate Member of the Institution of Electrical Engineers.

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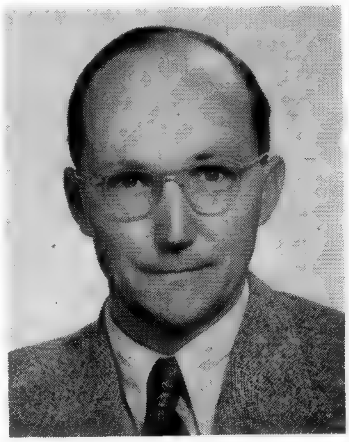

Theodor G. Grewe

Theodor Granz Grewe was born in Lippstadt, Germany, on July 20, 1916. He completed his studies at the Technical State College, Dortmund, in 1937 and the Technical University, Stuttgart, in 1948.

In 1938 and 1939, Mr. Grewe worked in the Research Institute of AEG, Berlin, on the development of acoustic altitude meters. During the second world war, he was employed in the torpedo testing installations at Eckenförde and Gotenhafen. On completion of his studies in 1948, he joined Standard Elektrik Lorenz A.G., Stuttgart, where he is a staff member in the transmission-technique laboratory.

Mr. Grewe reports in this issue on the theory of the negative-impedance two-wire repeater.

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Heinz Kürten

Heinz Kürten was born in Kempen, Germany, on November 26, 1925. He served as an aviator in the German Air Force from 1943 to 1945.

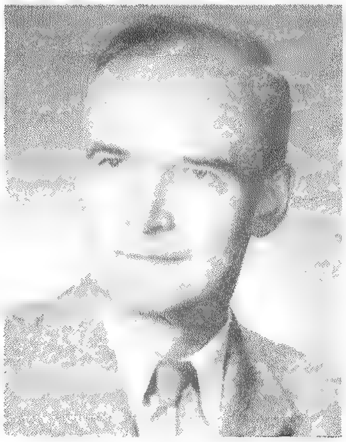
While working in industry, he also attended the Technical University at Aachen from 1949 to 1953. As a student, he visited Finland in 1951 and 1952.

On graduation as Diploma-Ingenieur, he joined Standard Elektrik Lorenz. He is now a project engineer in the export division. He is coauthor of a paper in this issue on Finnish crossbar toll offices.

Nikolaus Lewen was born in Wintrich, Germany, on February 17, 1910. After three years at Rheinische Elektrizitaets AG in Mannhein, he attended the Technical College at Bingen from 1933 to 1936.

He served as a telecommunication engineer for Siemens and Halske from 1936 to 1942. He then worked on the improvement of aircraft engines for Daimler-Benz in Stuttgart.

In 1946, Mr. Lewen jointed Standard Elektrik Lorenz as a circuit designer. He is coauthor of a paper on the Finnish crossbar toll offices in this issue.



Nikolaus Lewen

Edmond A. Thomas

Edmond A. Thomas was born in Chambery, France, on July 21st 1899. He graduated as an engineer from Institut Electrotechnique de Grénoble in 1922.

He joined the Compagnie Générale de Constructions Téléphoniques in 1936 and was assigned to the power-line carrier telephony division of the technical department. Mr. Thomas is the author of the paper in this issue on the carrier system for the Algerian power network.

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Kabelwerk (division), Stuttgart
Lorenz Werke (division), Stuttgart
Mix & Genest Werke (division), Stuttgart
Schaub Werk (division), Pforzheim

IRAN Standard Electric Iran A.G., Teheran

ITALY Fabbrica Apparecchiature per Comunicazioni Elettriche Standard S.p.A., Milan

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NORWAY Standard Telefon og Kabelfabrik A/S, Oslo

PORTUGAL Standard Eléctrica, S.A.R.L., Lisbon

PUERTO RICO Standard Electric Corporation of Puerto Rico, San Juan

SPAIN Standard Eléctrica, S.A., Madrid

SWEDEN Standard Radio & Telefon AB, Stockholm

SWITZERLAND Standard Téléphone et Radio S.A., Zurich

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Kolster-Brandes Limited, Sidcup
Standard Telecommunication Laboratories Limited, London

VENEZUELA Standard Telecommunications C.A., Caracas

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Companhia Telefônica Nacional, Curitiba and Pôrto Alegre

CHILE Compañía de Teléfonos de Chile, Santiago
Compañía Internacional de Radio, S.A., Santiago

CUBA Cuban American Telephone and Telegraph Company (50% owned), Havana
Cuban Telephone Company, Havana
Radio Corporation of Cuba, Havana

PERU Compañía Peruana de Teléfonos Limitada, Lima

PUERTO RICO Puerto Rico Telephone Company, San Juan
Radio Corporation of Puerto Rico, San Juan

SPAIN Compañía Radio Aérea Marítima Española, S.A., Madrid

UNITED KINGDOM International Marine Radio Company Limited, Croydon

ASSOCIATE LICENSEES FOR MANUFACTURE AND SALES

FRANCE Lignes Télégraphiques et Téléphoniques, Paris

ITALY Società Italiana Reti Telefoniche Interurbane, Milan

JAPAN Nippon Electric Company, Limited, Tokyo
Sumitomo Electric Industries, Limited, Osaka

SPAIN Marconi Española, S.A., Madrid

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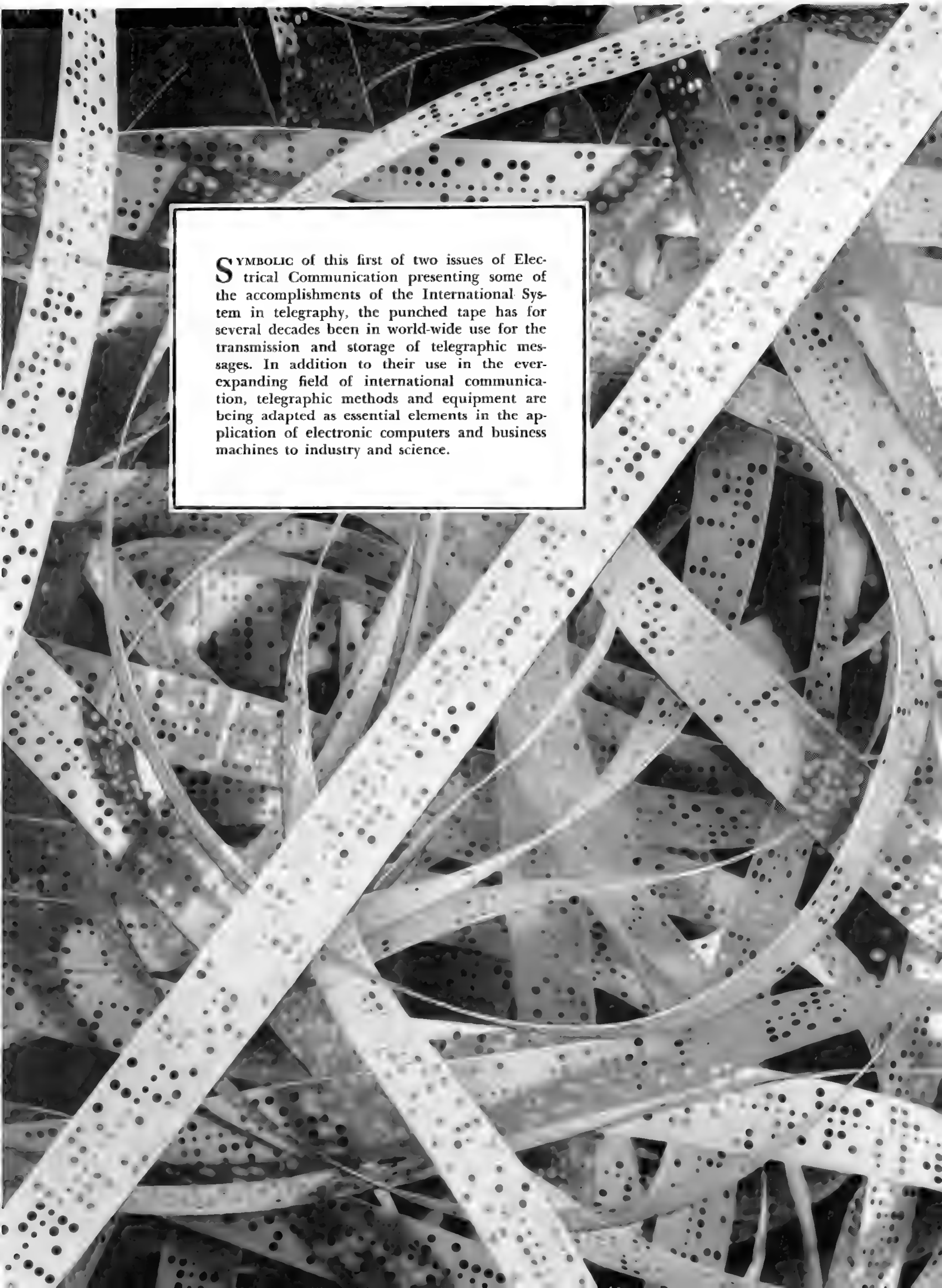

Symbolic of this first of two issues of Electrical Communication presenting some of the accomplishments of the International System in telegraphy, the punched tape has for several decades been in world-wide use for the transmission and storage of telegraphic messages. In addition to their use in the ever-expanding field of international communication, telegraphic methods and equipment are being adapted as essential elements in the application of electronic computers and business machines to industry and science.

STRAD—New Concept for Signal Transmission Reception, and Distribution

By E. P. G. WRIGHT

Standard Telecommunication Laboratories Limited; London, England

SUFFICIENT difference exists between electromechanical and electronic switching systems to provide ample material for a technical paper. However, few readers will have the time to study intricate details—they will probably be more concerned to consider what advantages an electronic system such as STRAD has over earlier systems—and as a consequence the paper commences with the basic advantage of the rapid processing that is now possible and the manner in which this speed can be employed to provide an economical arrangement of the various operations that need to be carried out.

It is explained that different requirements may need different arrangements; this state of affairs is similar to that experienced with electromechanical systems.

For readers who are familiar with the principles of electronic digital computers it will be evident that no new technique is essential.

Special procedures have been adopted to achieve a system reliability vastly greater than that existing in present computer systems for which a high degree of accuracy is necessary but for which a predetermined "down time" can be accepted. For communication systems, continuous service is necessary although a very small percentage of errors may be tolerable.

1. *Electronic Data Processing*

STRAD is an electronic version of a torn-tape system capable of receiving, storing, and retransmitting different forms of coded information. Arrangements can be made for push-button operator control, for automatic operation, or for a combination of these two methods. No electromechanical switches are employed, and the electronic devices introduce a reduction in annual charges both on the operating staff and on the maintenance staff required for the upkeep of telegraph instruments.

The speed of data processing possible with electronic components is so high in comparison with the customary telegraph line speeds that it is possible with time division to employ common high-speed cross-office circuits serving most, if not all, of the lines. This rapid processing of the data introduces economy and simplicity in a number of ways, one example being the handling of multi-address messages. In a torn-tape system it is customary to employ a tape factory to produce a separate tape for each direction of retransmission required. A single electronic record of the message is sufficient for any number of retransmissions. No delays are introduced before retransmissions can commence, and there is complete flexibility to proceed with the retransmissions as the outgoing circuits become available without any necessity for the retransmissions to be in phase with one another. The simplicity of the arrangement is emphasized by saying that the processing of each retransmission of a multi-address message is identical to that of a single-address message. In Figure 1 a message B is decoded at B_d and thereafter retransmitted B_r over a number of outgoing lines. Outgoing line 5

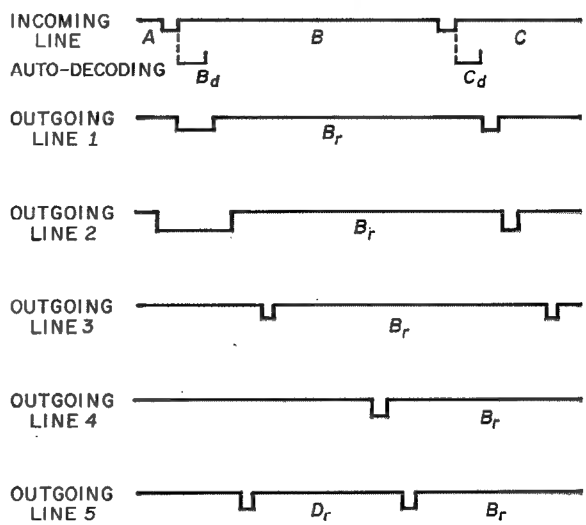

Figure 1—Retransmission of multi-address message using precedence control.

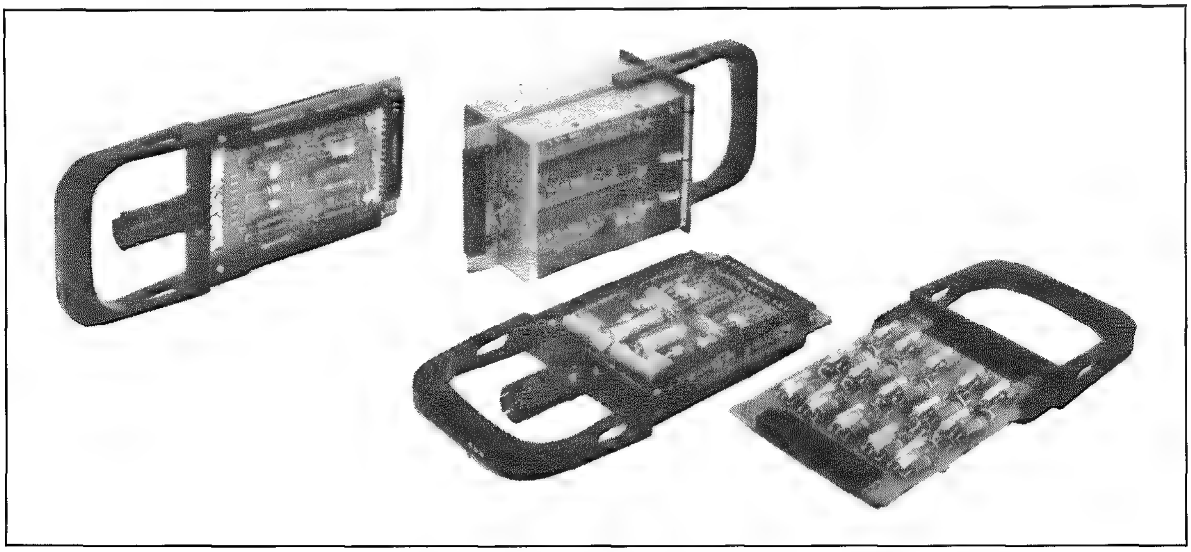

Figure 2—Typical plug-in units.

has to retransmit a message of higher precedence D_r before sending B_r.

Another striking example of the benefit obtained by rapid processing concerns the sorting of waiting messages of different precedence and chronological order. When an outgoing circuit becomes free at the completion of a retransmission, an examination is made of all the messages in storage to choose the next message in precedence and chronological order. It may happen that several outgoing lines are free at once, and in this case the message of highest precedence for all these lines is chosen first and then a similar message selection is made for the other waiting lines. The searching process involves only a few milliseconds because of the high speed.

There are many other ways in which the high processing speed is used to introduce economies. These will be described in detail later in the paper so that some consideration may first be given to reliability, faultfinding, and other aspects of maintenance.

2. *Reliability*

Some apprehension may arise on account of the extensive dislocation to the service that could result from a fault in one of the common circuits using time division, and this situation is catered for by a complete duplication of the common circuits with comprehensive cross-checking to isolate, disconnect, and alarm the section of the circuit that has developed a fault. The majority of the circuit units are of the plug-in type, so that a faulty unit can be identified, replaced, and tested *in situ* without interruption of the service. The importance of high-grade components and ample circuit margins needs no special emphasis.

It is a feature of STRAD that a relatively small number of different circuit units is employed to produce the various logical functions that need to be undertaken, and this arrangement goes a long way towards removing the necessity for the maintenance staff having to learn the operation of a variety of different circuits. The basic switching and processing operations are undertaken by triggers and gates, while the writing and reading of data into or out of the stores is undertaken by standardized amplifiers. It is not intended that any circuit-unit testing should be carried out on the racks, but only on the bench in conjunction with the appropriate testing devices. As a consequence the testing and repair of units can be undertaken under ideal conditions. Figure 2 shows examples of typical plug-in units.

It is appreciated that in such an electronic system, using pulses at a 50-kilocycle-per-second repetition rate, a serious fault liability can arise through the presence of dry joints due to imperfect soldering or due to wire fractures caused by excessive heat. To overcome this fault liabi-

lity, secondary solderless wrapped connections are employed, and the experience gained with this type of joint over a number of years has been entirely satisfactory. Special emphasis has been directed to this subject because the tracing of an intermittent failure of a pulse lasting only a few microseconds is necessarily elusive. Figure 3 shows secondary wrapped joints on a plug-in unit.

Mention has been made of the ability of a single link or highway circuit with time division to carry the traffic of a number of lines, and it is necessary to describe briefly what safeguards are employed to ensure that a serious interruption of the traffic is not caused by a component failure within the link.

The first step taken is to duplicate the link, then to make arrangements that both circuits operate in parallel, and then to check that they provide the same output. If the outputs should fail to check, an alarm is given, and it is then desirable that the failure should be analyzed and identified so that the faulty unit can be prevented from interfering with the output being supplied by the correct circuit. Figure 4 shows schematically this form of highway duplication. With such an arrangement the chance of disturbance is substantially reduced and the information provided to the maintenance staff enables a rapid replacement of the faulty unit to be made.

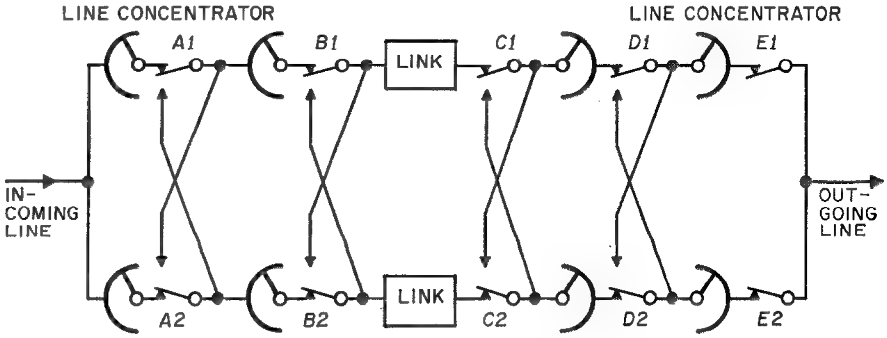

Figure 4—Highway duplication.

It is worth noting that complete equipment duplication is not restored at the time that the faulty unit is replaced because the link is associated with storage equipment in which information may have been stored incorrectly as a result of the fault and, therefore, when the link is returned to service there will be a possibility that the alarm will again be given as a consequence of the original fault. However, the difficulty is one that tends to be self-clearing because the contents of the stores are normally quickly refilled and at the end of a few seconds the links will pull into line if they are operating correctly. The faulty unit can be checked on the test bench with the appropriate testing equipment, but this check and repair of a faulty component is not a matter of any urgency; it is a much more urgent matter to replace in service the unit on which the failure occurred.

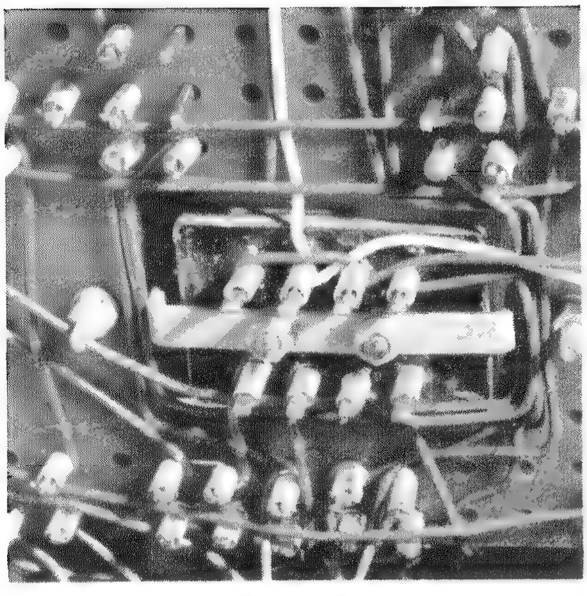
Figure 3—Secondary wrapped joints of a plug-in unit.

A simple example can be taken from the case of a group of incoming lines served by a highway with time division applied to allow the various lines to be sampled in sequence by a common timing circuit. A concentrator designed for this function causes the incoming signals to be reshaped and interleaved in sequence as output signals on the highway. The majority of component faults will appear as if the component is either short-circuited or disconnected. The

consequence on a time scale will be either that counting ceases or that it proceeds at double speed due to one trigger stage sticking in the operated condition. During the period between characters each circuit can be examined to see that there is no sticking, and, should any be found, the link concerned can be alarmed and steps taken to disconnect the multiplex output provided from the faulty concentrator. On the other hand, if the counting ceases, the concentrator automatically fails to give any output and it is therefore only necessary to give an alarm that there is a failure to check. In this case the

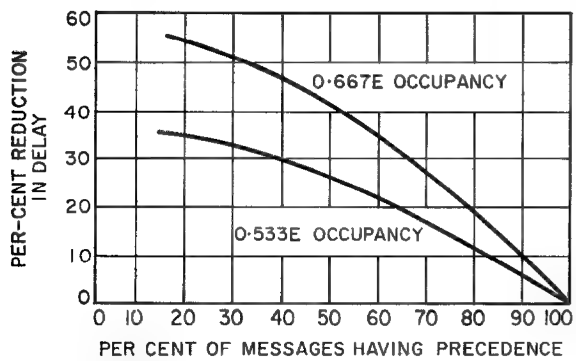

Figure 5—Effect of precedence on delay.

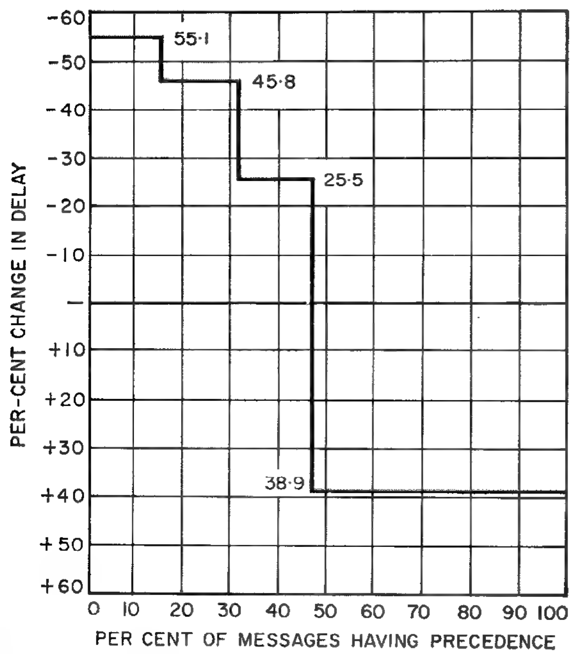

Figure 6—Relative effect on delay when using 4 precedence states.

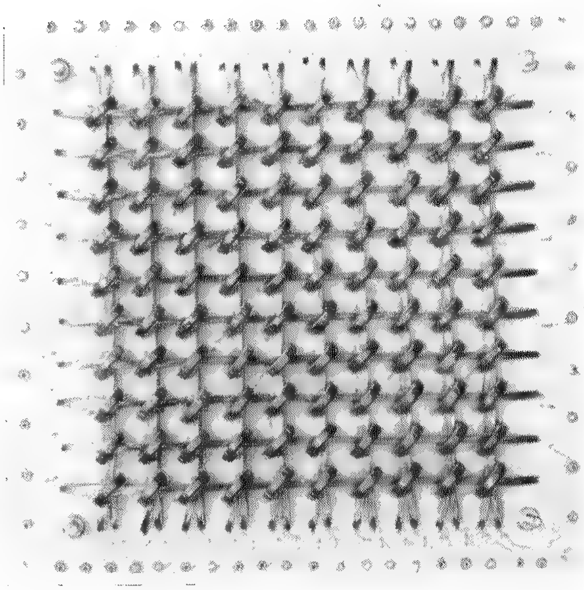

Figure 7—Ferrite access selector.

whole contents of the store are replaced in the period needed for the reception of two characters (say one-fifth of a second) so that the maintenance staff can determine without delay whether the faulty unit has been identified, replaced, and brought into synchronism with the other circuit.

3. *Storage*

A torn-tape system is normally employed to allow the incoming and outgoing circuits to be much more heavily loaded than would be possible with a direct switching system without storage in which the number of busy connections must be limited to avoid the additional operating that would otherwise be involved. Torn-tape systems are not usually expected to pass messages as quickly as direct switching systems. However, an electronic system with storage does not suffer from this disadvantage because the retransmission of a message can start a few seconds after the message commences to arrive. Heavy loading of the outgoing lines will inevitably introduce delays, but the use of different precedence categories will ensure that the urgent messages will be handled with a minimum of delay. With lightly loaded outgoing lines, the speed of handling precedence messages may be faster than with a direct switching system. Figure 5 shows the effect of precedence on delay, and Figure 6 shows the

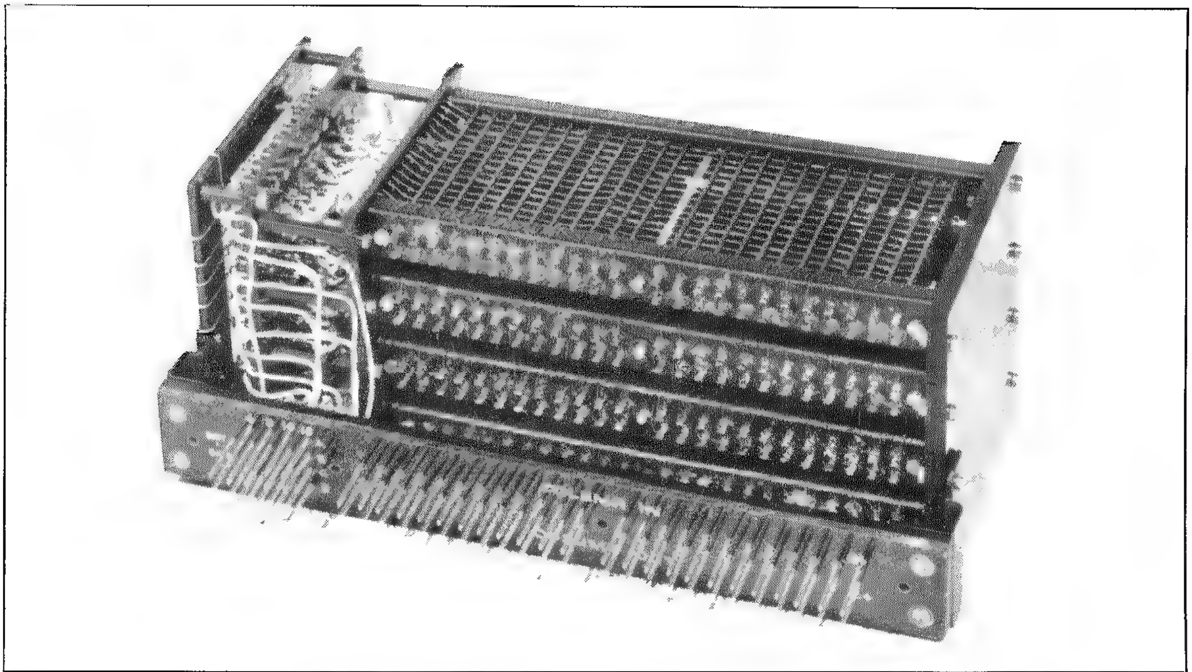

Figure 8—Ferrite store.

relative effect on delay when using four precedence states.

There are several interesting features in electronic switching systems, and for telegraph application perhaps the most important is the message store. There is a variety of different electronic devices available for the storage of telegraph characters, that is, shift registers, ferrite cores, magnetic drums, and magnetic tape. Figures 7 and 8 show a ferrite access selector and store,

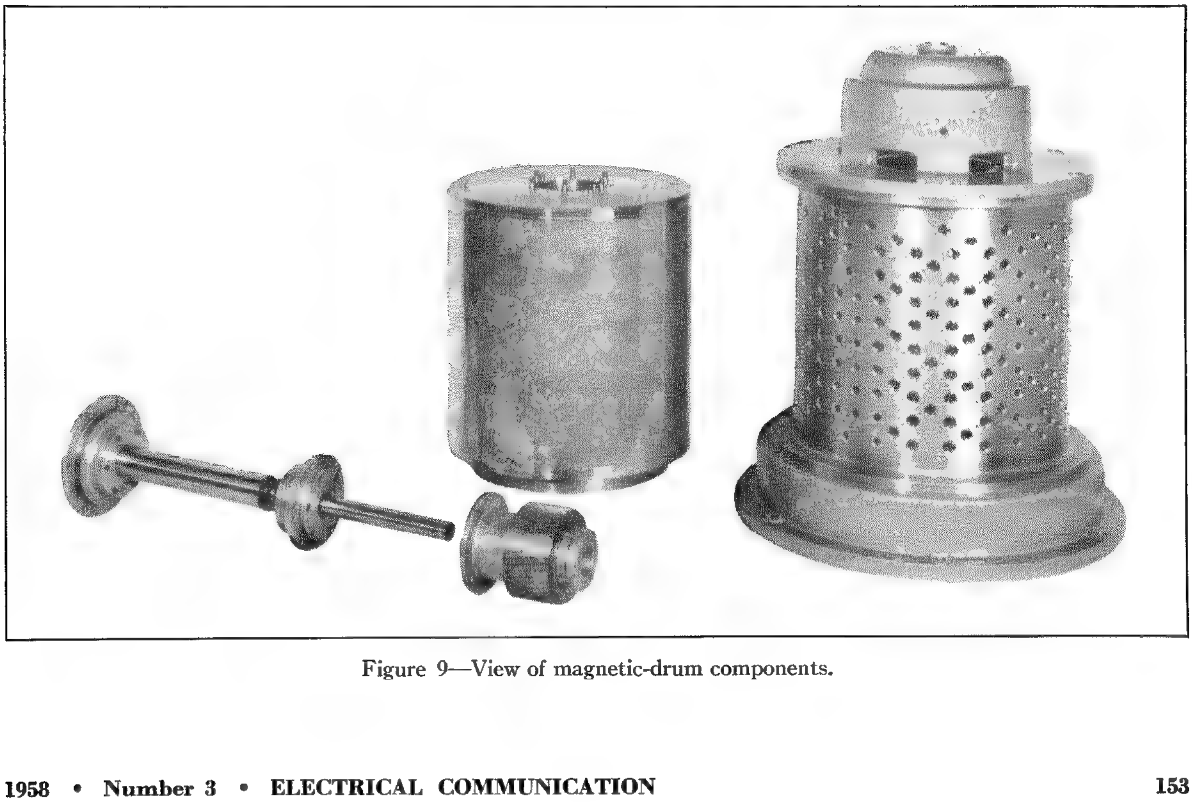

Figure 9—View of magnetic-drum components.

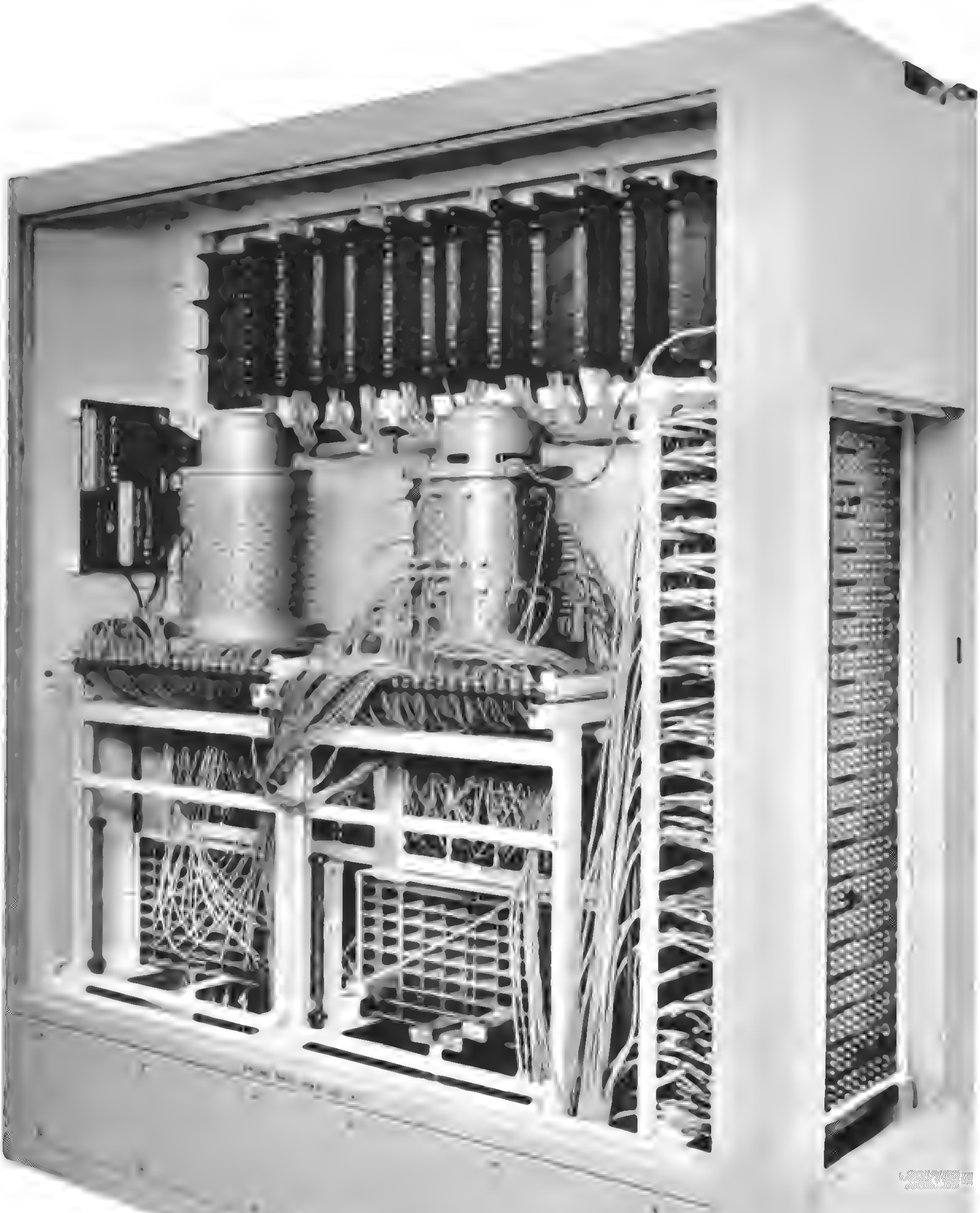

Figure 10—View of magnetic-drum console.

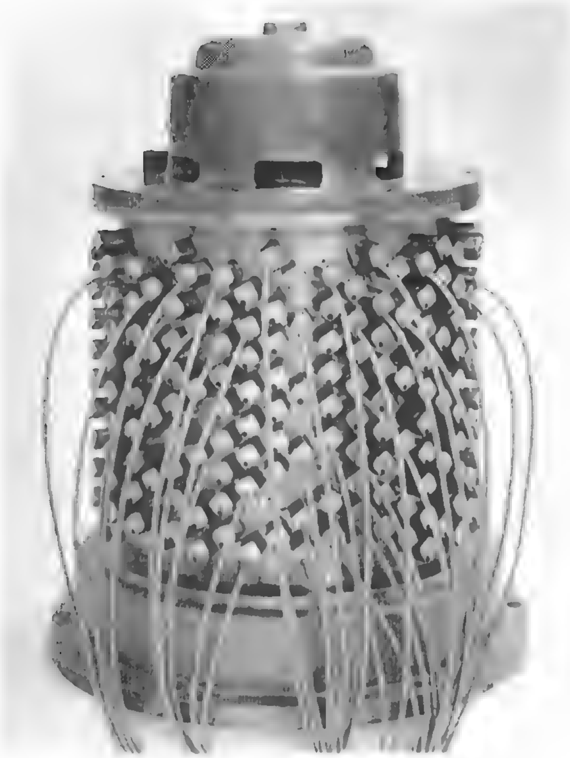

Figure 11—View of magnetic drum.

Figures 9, 10, and 11 show views of magnetic drums, and Figure 12 shows a large-capacity magnetic-tape machine. The suitability of these different devices depends on cost and access time. In the examples given above, the stores are arranged in order of decreasing cost and increasing access time. For applications in which a large amount of storage is needed it is advisable to use tape, while on the other hand, if the storage capacity needed is small, there are economies to be realized by employing ferrites because of the quick access that is available with them for writing or reading. Shift registers are too bulky and expensive to be used for the main store, but they are valuable for temporary storage for the following applications.

The shift register can be used as a time buffer to enable characters to be read at one speed and retransmitted at another speed. Such buffers can be used to record characters and transfer them onto the track of a magnetic drum or to withdraw the characters from the drum track and retransmit them to lines. The drum is rotating continuously, and it is necessary to employ a distributor to control the writing of successive characters in correct sequence on the track; this distributor can take the form of a shift register with an associated carrier (revolver) track.

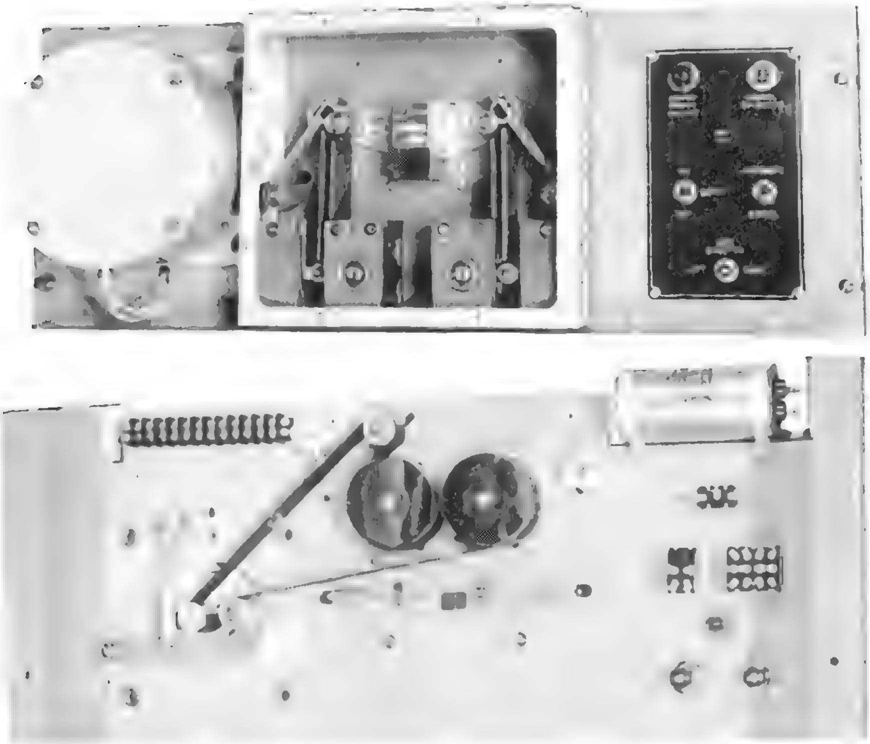

Figure 12—Large-capacity magnetic-tape machine; front view at top and rear below.

It is necessary to scan messages for various service indicators, and it is convenient to employ a shift register to scan each character in turn throughout a message.

The most efficient storage arrangement is that employing a single common store for all incoming and outgoing lines (Figure 13), but as all the lines may be in the process of receiving or transmitting at the same time there is clearly a necessity for quick access to the appropriate section of the common store to

ensure that each of the lines may be serviced sufficiently rapidly so that there is no danger of information being lost or delayed. The number of operations for the access selector can be materially reduced by writing and reading a whole character at a time as opposed to element by element.

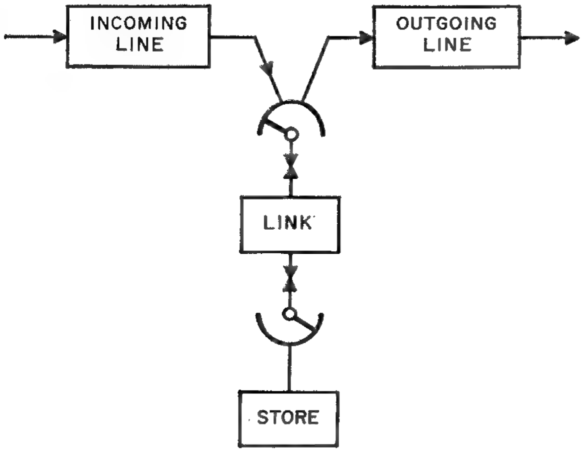

Figure 13—Block diagram for STRAD with common store.

Common storage is not only valuable because of its general availability but also because it allows retransmission to commence with negligible delay. Overall requirements for quick retransmission of messages may therefore demand a common store, and in general it seems likely that in such systems there will not be any need for a large store because the loading of the outgoing circuits cannot be allowed to be very high, otherwise delays will be introduced.

In other applications, usually with longer lines, there may be a need for higher line loadings to economize on line charges even though this may involve messages being delayed for a minute or two. This naturally necessitates more storage and, as a consequence, a cheaper form of storage such as a magnetic drum. However, the access time for a magnetic drum is necessarily greater and the use of common storage is more difficult to arrange. A compromise is obtained by assigning a drum track for each line and by transferring track loads of characters to the common store one track load at a time. As several hundred characters can be assembled on the line tracks, the number of transfers to the common store is greatly reduced and, furthermore, a complete track load can be transferred during a single revolution of the drum.

There are applications also in which it is necessary to allow for serious congestion when storage capacity is needed for several hours of continuous message reception. In such cases it is advantageous to employ the cheaper form of storage available with tape machines (Figure 14).

In addition, a tape machine can be employed as a buffer to absorb the incoming traffic at any time if it is desired to service the switching equipment. Although it might be expected that an interruption of this kind would result in a serious disruption in the traffic flow, this is by no means inevitable as the tape boxes associated with the outgoing lines will normally be holding a number of messages whose retransmission will continue until all the waiting messages have been sent. Furthermore, the cross-office speed through the switching equipment is many hundred times the line speed, so that, when the switching equipment is restored to service, the backlog will be taken up relatively quickly.

It is evident that the overall system requirements have a basic influence on the storage arrangements and that there is considerable flexibility for meeting widely differing traffic conditions.

4. *Traffic Calculations*

In an electronic switching system such as STRAD, in which one connecting link can usually carry all the traffic, the only circuits for which quantities have to be calculated are the stores. However, the stores may represent a considerable percentage of the initial cost of the equipment

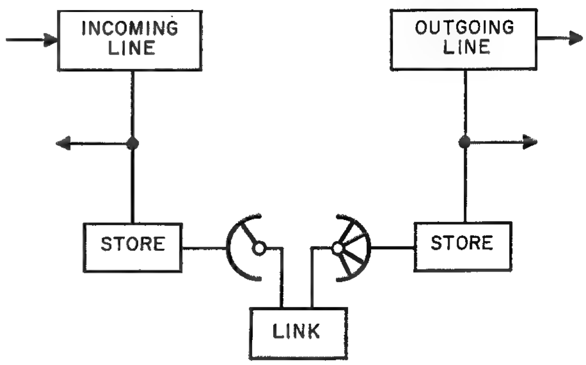

Figure 14—Block diagram for STRAD with line stores.

and as a consequence the quantity calculations merit careful examination.

If the storage is provided on a line basis rather than on a common storage basis, there is no need for calculation because the capacity required can be stated directly.

The conditions are different in the cases for which common storage is provided. If, for example, drum tracks are used, and if the average message length is shorter than the drum track, then the traffic in erlangs can be taken as the product of the number of messages multiplied by the holding time of the store expressed in hours. Normally this holding time will be the delay before the message is retransmitted, and Figure 15 shows this delay in relation to the message

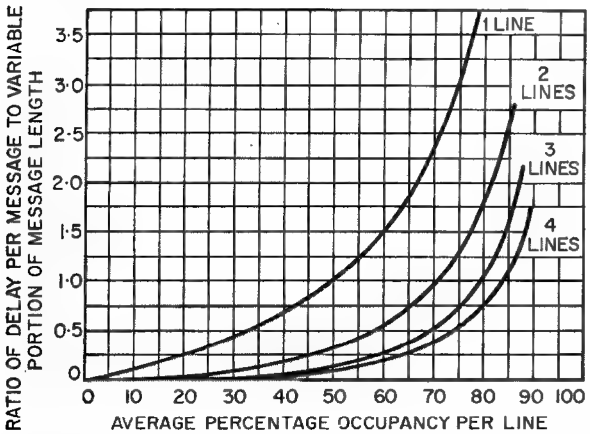

Figure 15—Average delay per message as a multiple of average message length in respect to occupancy for different line groups.

length and the line occupancy. The number of stores needed for this traffic at any specified loss can then be calculated from available probability curves.

Similarly, if the average message length necessitates 2 drum tracks, the traffic will be approximately 2 multiplied by the number of messages, multiplied by the holding time of each track. It is necessary to point out that, if the average message length is exactly twice the capacity of a drum track, it is not reasonable to assume that the average number of tracks required will be 2; there will be many messages of less than average length that still require 2 tracks and many messages of more than average length that require at least 3 tracks. The correct quantity is, therefore, likely to be nearer $2\frac{1}{2}$ tracks per message than 2. An exact answer can be obtained only by examining the distribution of holding times during the busy hour.

When common storage is provided by ferrite cores the calculation is more involved because

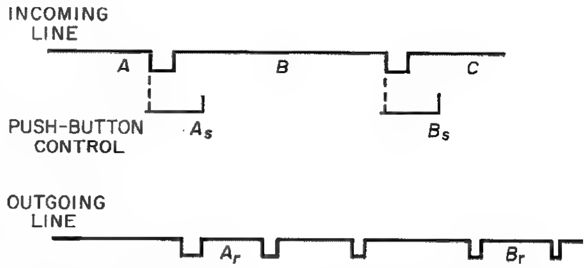

Figure 16—Message retransmission with push-button control of incoming messages.

the size of the store section is not controlled mechanically and its optimum value has to be calculated. The store traffic will approximate to the average message length (in words) divided by the store section capacity (in words), plus a half, and multiplied by the holding time of the store section. In the earlier example the holding time is substantially the average delay before retransmission commences. In the case of the ferrite store the holding time is the average delay plus the time taken to empty the store. If the cadence speed for filling the store is lower than the autocadence speed for emptying the store, then the holding time for each store section used for the average message will be different and a separate calculation will be necessary for each. Figure 16 shows a message B received at low cadence speed, directed by push-button control at B_s and retransmitted at B_r at autocadence speed.

Having calculated the traffic on the lines indicated above, the number of stores can be determined for any appropriate loss probability. However, it will sometimes happen that no free store is available when needed, and there is no simple way of indicating to the sender that his message has not been wholly recorded. This state of affairs cannot be satisfactorily overcome by reducing the loss probability to 1 in 100 000 or 1 in 1 000 000 because the congestion may be introduced by abnormal delays that are much more severe than the average traffic flow on which the

loss calculations are based. Relief can be obtained, however, because the percentage occupancy of the whole common store can be continuously monitored by automatic means, and, as soon as some predetermined value is reached, overflow circuits can be introduced to transfer some of the waiting messages to tape storage. Figure 17 shows a small-capacity low-speed magnetic-tape machine. These messages can be returned to the store as soon as the occupancy falls to some second predetermined value. The messages temporarily withdrawn in this way may be subjected to some extra delay, and in consequence it is clearly desirable that the messages selected should be of low precedence.

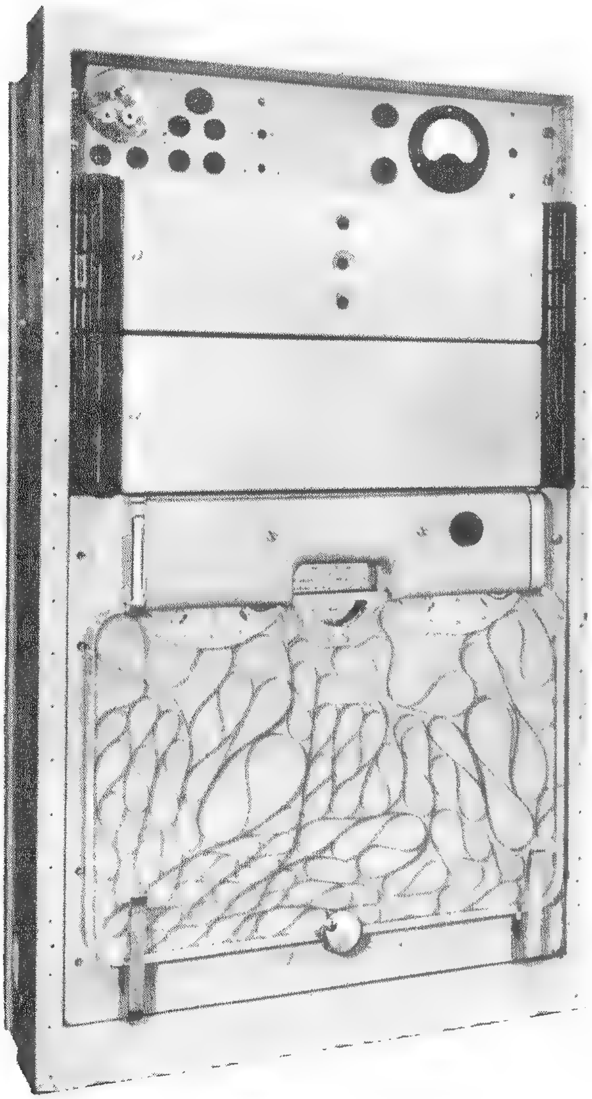

Figure 17—Small-capacity magnetic-tape machine.

In the foregoing description the average delay before retransmission commences has been mentioned. This average delay has to be determined from data providing the occupancy of the outgoing lines and the average holding time of the messages (see Figure 15). In the case of multi-address messages, only one message needs to be counted, but the holding time will be influenced by the number of retransmissions that have to be made. The delay times in all cases increase steeply as the outgoing line occupancy approaches 100 per cent, and, if the traffic offered to a line exceeds 100 per cent during a busy hour, there will inevitably be some overflow to the following hour. This fact emphasizes that from the point of view of the store the busy hour is not necessarily the hour in which most messages occur but the hour in which most messages are awaiting retransmission. The busy-hour traffic will vary from the mean each day, and it is unfortunately necessary to collect reliable statistics of these variations if an accurate estimate is to be made of the storage requirements. The advantage of common storage is striking in this connection because, instead of having to collect statistics about each line, it may be sufficient to take the overall traffic and to estimate the average message delay while awaiting retransmission for all lines because it is of little importance which line has the greatest delay if the distribution pattern is known.

5. *Electronic Switching*

The electronic switching equipment employed is basically the connecting link between the lines and store. In the cases in which a common store is provided, a switch on one side of the connecting link gives access to the different parts of the store, while on the other side of the link there is another switch giving access to the different incoming and outgoing lines (see Figure 13). If magnetic drums are employed, then the link provides a connection for a period of about 100 milliseconds for the transfer of a track load of data and for the various control operations that are carried out at the same time. It may be calculated that it takes approximately 40 seconds at line speed to fill or empty a track, and it is, therefore, evident that

even with heavy line loading a single link will be able to cater for a number of lines.

With the use of ferrite storage the arrangements are rather different because the link is provided with a character store, and it is, therefore, possible to draw a distinction between the use of the connecting link to transfer element information from the lines to the link store or vice versa and its use for transferring complete characters from the link store to the common store and vice versa. The link is set up several times per millisecond for the transfer of element information, and approximately one operation in seven includes the transfer of a complete character to or from the common store. As before, a single link circuit can handle the traffic for a considerable number of lines.

The link arrangement is different for the case in which larger stores are provided per line because it is then necessary to connect one side of the link to the incoming lines and the other side to the outgoing lines. As, however, the message may have several addresses, it is arranged that the switch on the outgoing side can make connection simutaneously to any combination of these lines.

The provision of supervisory and control facilities in the link circuit to recognize indicators and check serial numbers represents an economic arrangement because a single circuit is serving so many lines. In the cases in which a common store is used, the link is also an economic arrangement to examine and compare data because here again a single circuit is serving a number of lines (see Figure 18).

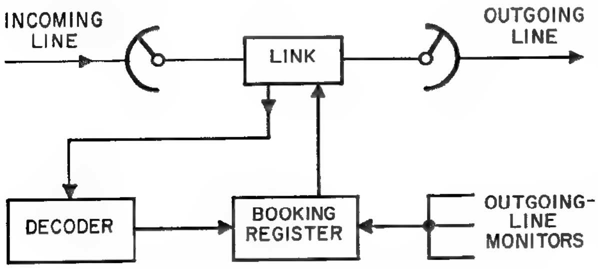

Figure 18—Block diagram of message booking and retransmission.

The means by which the connecting link makes access to the different lines is interesting to examine in some detail because there is a variation between the different cases. When there is a store associated with each line, the link switch gives access directly to these stores and in consequence there is one position for each line. On the other hand, in the case of the ferrite stores it is possible to employ space-to-time conversion in the form of a concentrator, and as a consequence the link switch only needs to give access to these concentrator outlets or inlets and not to the lines. In the case of the incoming circuits some 16 or 20 lines are grouped on a single concentrator which has an outlet providing in multiplex form the mark/space condition of these lines that the link circuit receives and distributes to the different character stores that it contains. Somewhat similarly on the outgoing side the multiplex lead is informed of the mark/space-element conditions to be transmitted by each of the lines in the group, and it is one function of the outgoing concentrator to distribute these signals to the different lines in the group.

The use of these concentrators, therefore, serves to simplify and reduce the size of the switches associated with the link and also to reduce the wiring involved in making connection to each individual line. In addition, the concentrators are used to make an economy in the hardware, that is, by providing time scales both for the examination of the lines to read the incoming data and for the corresponding process involved in retransmitting the data. To achieve this purpose the incoming-line concentrator employs a small matrix in which each row corresponds to a different incoming line. These rows are read continuously in a fixed sequence, and a number of column circuits together act as a binary counter, so that when required a unit can be added to the number read and the new value can be rewritten in the matrix. This counting operation is controlled by data also inserted in the row information as a consequence of the start element being received for each character in turn. Once this counting process is commenced by the start element it is continued automatically until the link circuit signals that it has received the complete character. The use of an electronic time scale to provide the necessary examination pulses calls for no description, but it may be explained that as a consequence of the examination the condition of the line is also recorded in a storage element in the row associated with the line examined; this information is held available until

the link circuit is in a position to ask for it. The introduction of this element storage in the matrix associated with the line concentrator allows the latter to operate at a repetition rate that is sufficiently high to reduce the ratchet error (introduced by the sequential examination of the lines) to a small fraction of a millisecond, which is of minor consequence. On the other hand, the effective repetition rate of the link can be reduced to that necessary for collecting and distributing the elements of the different characters received from the incoming lines. The repetition rate used for the link may be increased with the number of concentrators connected, but for centers having fewer than 120 lines this repetition rate is still much lower than that of the concentrators. This lower rate provides an advantage in allowing more time for the entry and withdrawal of characters to and from the common store.

The concentrator principle is also of interest for the control of a number of way stations on an omnibus line. One way of operating such an arrangement includes interrogating the stations in turn to ascertain whether they have a message waiting for transmission. The storage matrix can be used in conjunction with a set of column circuits to act as a distributor so that each station is signalled in turn. It will be appreciated that it is also necessary to design this distribution arrangement so that each new interrogation commences at the point at which the previous one ceased. It is a simple and economical arrangement to maintain the data in a row of a ferrite matrix and to employ common column circuits for processing this information at the proper times.

6. *Message Length*

There is no fundamental restriction on the length of message that can be accepted, but there are practical reasons for fixing a limit. For example, a long message of low precedence may hold up the retransmission of a more urgent message unless facilities are provided for interrupting messages in process of retransmission.

Furthermore, in the event of a failure to recognize the end-of-message indicator, due to distortion, an alarm can be introduced when reception exceeds the maximum length; for accounting and statistical purposes, facilities can also be provided for measuring the number of words in a message.

7. *Flexibility*

There is no restriction on the character combinations that may be chosen for the start-of-message, the end-of-message, or the end-of-routing indicators and similarly there is no restriction on the number of letters or on the combination of letters and figures used for the routing indicators. A booking register is used to record particulars of messages received and to provide translation facilities for relating the routing indicators with the lines or trunks with which they are associated. The booking register is connected to a link via a decoder, as shown in Figure 18, and is used for initiating the retransmission of messages under the control of the available outgoing lines.

The translator circuit in the booking register is designed in such a way that the routing indicator combinations and their line associations can be changed from time to time without any equipment or wiring rearrangment. This is not intended as a saving in material but rather as a means of avoiding any instability in service that could result from making physical modfications in the field. It will be appreciated that certain advantages can be obtained by choosing routing indicators that are abbreviations of the names of the organizations concerned because this will lead to simpler and shorter messages. On the other hand, there is some advantage in choosing routing indicators so that the first few characters identify the switching centers, and thus the translator has a much simpler function to perform. If codes are being selected to suit the switching system, it is advisable to choose a series that permits error indication by parity checking.

Arrangements can be made for STRAD to remove from message retransmissions the routing indicators corresponding to stations for which retransmission has been provided elsewhere. This is a rather-more-complex arrangement than causing the translators to obey a program that orders that certain predetermined indicators are disregarded on messages from other switching centers. In both cases facilities are provided for changing the program to admit alternative routes introduced as a consequence of congestion.

Arrangements can also be made for the automatic "erasure" of surplus incoming-channel identification and message numbers from the retransmitted message. The facility might be useful in large networks employing a number of interconnected switching centers to eliminate the cumulative build-up of the original message preamble.

For networks employing several precedence categories, a "break-in" facility can be provided to ensure that a message in course of retransmission can be interrupted to allow one of higher precedence to be cleared. This facility is probably only worth consideration for networks handling long messages, and it is evident that a message of the highest precedence should not interrupt a message of equal importance.

These different facilities are made economically practicable because they can be applied to the messages as they traverse the link at a speed of 50 kilobauds (approximately 83 000 words per minute with the 5-unit teleprinter code).

The line terminal conditions for STRAD are unrestricted. The system is designed for duplex channelling, both lines and trunks having independent send and receive channels. The line terminations may employ either manual keyboard transmission or autotransmission. The lines can be associated with one or a number of stations. The trunks can terminate in torn-tape, push-button, or automatic switching systems. For radio channels it is likely that the problems of propagation, frequency changing, and interference will make it desirable to employ reperforators and autotransmission.

8. *Monitoring and Supervision*

On a push-button system the operator can either book the direction for a message when it appears in the text or she can wait until she has checked that the message is correctly terminated before booking. With automatic switching it is arranged that any message with an unidentified precedence or routing indicator is passed to the supervisor who can reinsert it into the system with push-button indications of the action to be taken.

In addition a record is made of the essential parts of the preamble of all incoming messages; a separate record of outgoing messages provides a cross reference of channel-identification serial numbers. A record can also be produced on request of all waiting messages.

Facilities can be provided for periodic line testing so that in the absence of a message for some predetermined period a signal is generated asking for a test message. A copy of this request is sent to the supervisor. If no response is forthcoming within a reasonable period, an alarm is given so that the line can be tested.

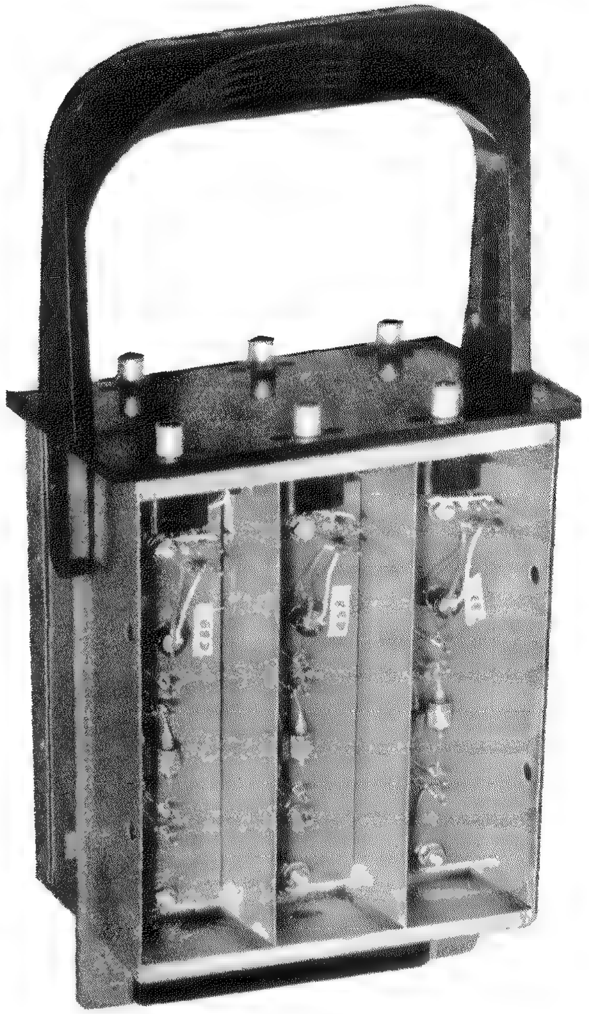

Figure 19—Typical switching unit.

9. *Hardware*

It would have been possible to present this paper with the main emphasis applied to the components. It is true that the components are of extreme importance in respect to performance, but STRAD does not depend on a particular range

of components and the system could employ hard tubes, gas tubes, or transistors for the logical functions. It has been mentioned elsewhere that different types of storage devices may be used. It is of interest to state that in all cases symmetrical transistors are employed for switching purposes, and Figure 19 shows a typical switching unit.

10. Conclusions

It may well be argued that no worthwhile conclusions can be reached about a new switching system until field experience has demonstrated practical results. Nevertheless it is pertinent to observe that most of the electronic technique used in STRAD has already been used elsewhere and that this part of the development is not therefore new but in line with progress on digital computers and other data-processing systems.

A major question that arises with any new system employing entirely different components is whether it represents merely a stepping-stone in a long-term development or whether it constitutes a significant advance justifying immediate adoption. In this connection it should be pointed out that contemporary component development is very likely to provide either better or cheaper components that can be incorporated as they become available. It is difficult to forecast whether entirely new components will render the design of the system obsolete within a few years, but it will be difficult to justify any higher cost for components capable of operating at a higher speed because in most respects the speed of existing components is adequate.

11. Acknowledgments

The author acknowledges the assistance of many colleagues in Standard Telecommunication Laboratories Limited and Standard Telephones and Cables Limited, and in particular that of Mrs. E. Loverseed.

12. References

12.1 E. P. G. Wright and J. Rice," Probability Studies Applied to Telecommunication Systems with Storage," Proceedings of the First International Congress on the Application of the Theory of Probability in Telephone Engineering and Administration, Copenhagen, Denmark; June, 1955: also, *Electrical Communication*, volume 33, pages 308–321; December, 1956.

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Basic Considerations in Calculating Storage for an Electronic Telegraph Switching Center*

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OUTSTANDING among the features of an electronic switching center employing time sharing is that there is little necessity to calculate switch quantities to satisfy traffic requirements. A single highway can be used for a number of channels, and the determination of how many channels is partly influenced by traffic considerations but predominantly by the maximum pulse repetition frequency that is acceptable. However, the amount of storage to be provided for messages awaiting retransmission does constitute a calculation that is comparable with the switch quantity calculations undertaken to determine the number of registers necessary for a telephone exchange. Accurate storage calculations are a matter of basic importance both in relation to the initial cost and to the annual charges. For a system in which the storage requirements are modest, because the delays before retransmission are short, the cost of the store may represent less than half that of the whole center; but if the delays are long and there is a requirement to store hundreds or thousands of messages, then the majority of the cost of the center will lie with the storage equipment. Extra operating expenses will be involved in relieving the store manually if and when the store capacity proves to be inadequate.

1. *Switch Calculations for Electronic Systems*

To derive the amount of traffic to be stored, design curves[1] have been established from the number and occupancy of the outlets from the center, and from the time distribution of the message lengths. It is of interest to observe that the average message delay is a function of the variable part of the message length, and that the average delay can be appreciably reduced if there is an effective maximum message length of approximately 20 minutes.

Most design curves imply that with a circuit loading of 100 per cent the average delay will reach infinity, and before examining the calculations of storage capacity it is desirable to pay some attention to the question of outlets that are offered messages totalling 60 minutes in an hour.

2. *Busy-Hour Traffic*

2.1 Message Accumulation

It is evident that, if 60 minutes of messages to be retransmitted are presented during the busy hour, the number of message minutes awaiting retransmission at the end of this busy-hour period (MMAXE) cannot be less, but may be greater, than the number of message minutes awaiting retransmission at the initiation of this period (MMAXI). If there is a succession of such busy-hour periods following one another, the final MMAXE will be determined by the busy hour in which the distribution of messages is most irregular and none of the other hours will have any effect on this final MMAXE.

The irregularity of the incidence of messages has an unfavourable influence if at any time during the hour there remain no messages available for retransmission and as a consequence line time is wasted. As the number of hours is increased, the greater is likely to be the irregularity in any one hour. Conversely, the greater MMAXI becomes, then the greater is the likelihood of the traffic in an hour maintaining continuous retransmission. In fact, if MMAXI reaches 60 (minutes), it is no longer possible to experience a break in retransmission however irregularly the messages appear.

* Presented at the International Teletraffic Congress (Second International Congress on the Application of the Theory of Probability in Telephone Engineering), The Hague, Netherlands; July, 1958.

[1] E. P. G. Wright and J. Rice, "Probability Studies Applied to Telecommunication Systems with Storage," Proceedings of the First International Congress on the Application of the Theory of Probability in Telephone Engineering and Administration, Copenhagen, Denmark; June, 1955: also, *Electrical Communication*, volume 33, pages 308–321; December, 1956.

2.2 100-Per-Cent Busy-Hour Occupancy

It is necessary to consider what is meant by 100-per-cent busy-hour occupancy. It must certainly include the case in which 60 minutes of retransmission are offered during the busy hour, but should it not also include the case in which 24×60 minutes of retransmission are offered during the day? This broader conception is again capable of further analysis:

A. A smooth intensity of 100-per-cent occupancy.

B. A random condition in which peaks occur at certain specified periods of the day.

C. A random condition in which peaks occur at random periods of the day.

A smooth intensity over a number of successive hours must be accepted under the term 100-per-cent occupancy.

If peaks occur at certain specified periods of the day, only these periods should be considered in calculating the busy-hour traffic, and the traffic offered during these periods must exceed 100-per-cent occupancy.

If the peaks occur at random times, there are two precedents to consider:

A. The recommendations made by the Comité Consultatif International Télégraphique et Téléphonique for semi-automatic traffic; in principle these recommendations would accept the condition as 100-per-cent occupancy.

B. The procedure followed by certain administrations according to which only the busiest hour in each of a succession of days is considered. With this interpretation the traffic condition would exceed 100-per-cent occupancy.

TABLE 1

Throwdown Calculations

MMAXI	MMAXE			Message Delay in Minutes (Average over hour)		
	Minimum	Average	Maximum	Minimum	Average	Maximum
0	0	12·19	21·74	5·02	8·94	13·89
3	3	12·44	21·74	5·02	9·04	13·89
6	6	12·86	21·74	5·03	9·77	15·70
9	9	13·22	21·74	5·23	11·09	18·71
12	12	15·59	21·74	5·43	12·70	21·71
15	15	16·76	21·74	6·77	14·51	24·71
18	18	18·67	21·74	8·57	17·26	27·71
21	21	21·06	21·74	10·37	19·94	30·71
24	24	24·00	24·00	13·17	22·89	33·71

Average message length = 4 minutes
Maximum message length = 20 minutes
Minimum message length = 48 seconds
Message length distribution = exponential
Number of hours observed = 12
Messages per hour = invariably 15

TABLE 2

Percentage Distribution of Individual Message Delays in Throwdown Described in Table 1

Delay in Minutes	Per Cent	Delay in Minutes	Per Cent
0	17·75	13–14	2·5
0–1	2	14–15	2
1–2	5·5	15–16	3·25
2–3	7	16–17	3·75
3–4	9·5	17–18	3·25
4–5	3·25	18–19	6·5
5–6	5·5	19–20	2
6–7	2	20–21	2
7–8	1·75	21–22	1·75
8–9	7	22–23	2
9–10	1	23–24	0·5
10–11	2	24–25	0
11–12	2	25–26	1
12–13	3·25		

For the purpose of determining the store capacity, it is impractical to take the average traffic over a period but essential to consider the peak conditions within this period. These peak conditions are not caused solely by the messages offered during the most intense 60-minute period because the storage requirements are likely to be influenced to an important extent by the MMAXI value, as will be seen from Tables 1 and 3.

The MMAXI value is dependent on the in-

tensity and irregularity of the traffic during one or a number of preceding hours, and as a consequence storage requirements should not be based on the estimated delay for the traffic in any single busy hour but rather on the maximum number of message minutes likely to accumulate in a sequence of hours.

The most general case arises when the average of the busiest hour of a number of consecutive working days in the busy season equals 100-per-cent occupancy. In such a case there will be many hours above (and below) the average. There is likely to be a small chance, say 2 per cent, of the load reaching 150 per cent in one hour. It is evident that with 150-per-cent loading the MMAXE cannot be less than 30; the likelihood that there will occur pauses, during which there are no messages to transmit, must be less than with 100-per-cent loading, and it therefore seems unlikely that the MMAXE will often exceed 40 or 45. If two such hours follow one another, the MMAXE is likely to reach approximately 75.

3. *Throwdown*

It is a tedious undertaking to make a throwdown under the above conditions because the maximum MMAXE and delays will arise only in a small percentage of the hours examined, and yet all must be calculated to find the overall average. The results quoted in Tables 1 and 3 relate to throwdowns with an invariable loading of 100 per cent and may be considered as typical examples of excessive traffic hours from a group whose average is only approximately 75-per-cent loading. A number of throwdowns with 100-per-cent loading have been made, and from these throwdowns two sets of results are used to illustrate the typical accumulation of message minutes awaiting retransmission as a consequence of 100-per-cent loading during one or more busy hours.

3.1 4-Minute Messages

It should be stated for Table 1 that the busy hour that gave the maximum MMAXE figures also provided the minimum message-delay figures, while the hour responsible for the minimum MMAXE figures corresponded to the hour providing the maximum message-delay figures above the MMAXI figure of 3. In the first case the bulk of the messages occurred late in the hour, whereas in the second case the bulk occurred early in the hour.

TABLE 3

THROWDOWN CALCULATIONS

MMAXI	MMAXE			Message Delay in Minutes (Average over hour)		
	Minimum	Average	Maximum	Minimum	Average	Maximum
0	3·76	5·36	9·18	2·80	3·72	4·37
3	3·76	5·36	9·18	3·17	4·41	6·10
6	6·00	6·53	9·18	3·54	5·67	8·75
9	9·00	9·02	9·18	6·54	8·67	11·75
12	12·00	12·00	12·00	9·54	11·67	14·75

Average message length = 40 seconds
Maximum message length = 20 minutes
Minimum message length = 5 seconds
Message length distribution = exponential
Number of hours observed = 6
Messages per hour = invariably 90

TABLE 4

PERCENTAGE DISTRIBUTION OF INDIVIDUAL MESSAGE DELAYS IN THROWDOWN DESCRIBED IN TABLE 3

Delay in Minutes	Per Cent	Delay in Minutes	Per Cent
0	3	7–8	9
0–1	8	8–9	10·5
1–2	11·5	9–10	3·5
2–3	15	10–11	0·2
3–4	12	11–12	0·2
4–5	9	12–13	0·4
5–6	10	13–14	0·2
6–7	7·5		

In Table 2 it is to be noted that with 60 minutes of message retransmission offered in each hour (100-per-cent loading) 17.75 per cent of the messages were retransmitted without delay. The increased delay apparent around 15–19 minutes is a consequence of the longest message present

in each of the hours having affected several messages immediately following. This distribution is made assuming an MMAXI figure of 3; it can be seen from Table 1 that the delays would not have been seriously altered if a figure of 0 or 6 had been taken.

If the storage requirements are associated with the MMAXE figures, it would seem that for a single hour it would be sensible to assume a figure of about 16 minutes, for two successive hours perhaps 19 minutes, and for more successive hours perhaps 21 minutes. It is evident that a 12-hour throwdown is insufficient to indicate the worst irregularity that can occur, but it is probably unnecessary to cater for an irregularity that rarely manifests itself. A further point for consideration concerns the probability of similar peaks occurring on the different outlet groups simultaneously; this factor is of interest, particularly if the store is common to many outlet groups.

3.2 40-Second Messages

For Table 3 it should be stated that the busy hour that gave the maximum MMAXE figures also produced the minimum message-delay figures.

The delays shown in Table 4 are consequent on an MMAXI figure of 3.

4. *Influence of Message Length and Number of Traffic Sources on Delay*

4.1 Message Length

It can be seen from examination of Tables 1 through 4 that in general the shorter average message length will result in shorter message delays; this is apparent with all values of MMAXI for the maximum delays but only with the low values of MMAXI for the minimum and average delays. For the MMAXE figures there is a 2:1 ratio for the maximum but little difference for the minimum. It is perhaps surprising that the percentage of messages not delayed is much greater with the longer holding time. As the ratio of the holding times is 6:1, it is evident that no linear relation exists. If the average message holding time is increased to an impractical figure, such as 30 minutes, it is difficult to imagine the MMAXE going beyond 30.

4.2 Number of Traffic Sources

It is logical to assume that the greater the number of potential sources the less will be the chance of several messages coming from the same source within a busy hour. The examination of the subject is complex because, even if it can be established that, say, 25 or more sources are involved, it is unlikely that they will each offer the same percentage of the traffic. If several messages are habitually offered from the same source, then the chance that these messages will delay one another is eliminated because any such delay will occur at the source and not at the switching center. If there is only one source there will be no delay. This condition is likely to be somewhat artificial and one that will have little influence on the provision of storage equipment. Also it must mean that 100-per-cent loading from one source is not the average of several hours but the loading of either one hour or each of a number of hours.

4.3 Throwdown for 20, 10, and 5 Sources

The throwdown described in Table 1 has been repeated with 20, 10, and 5 sources, and the results appear in Table 5. It should be added that the number of sources had no influence on the minimum, average, and maximum MMAXE values.

The change in message-length distribution is of interest; with only 5 sources, about 25 per cent of the delays exceeding 19 seconds are reduced below this duration while another 20 per cent are reduced but not sufficiently to fall below 19 seconds. Furthermore, the percentage of messages with no delay increases to 20 per cent by virtue of certain follow-on messages that occur as the outgoing circuit becomes free.

It may be observed that the reductions in average delay are not varied by the MMAXI values. If the average delay with infinite sources is taken from Table 1 as 9 minutes and a curve is drawn of the average delays from Table 5 to terminate in no delay with one source, it will be found that the curve approximates to:

$$d_n = d_\infty - \frac{d_\infty}{n}$$

where d_n is the delay from n sources when the delay with infinite sources is d_∞.

This equation appears logical for examples in which most of the delayed messages have to wait for only one previous message to be completed. In such examples, with messages derived from many sources, the messages may be assigned to one or other of two empirical sources, and it will be seen that half of the message delays will then occur at the source rather than at the switching center. For the cases in which a message has to await the completion of more than one previous message, it will still be found that, when there are only two sources, half of the delayed messages will be delayed at the source.

It may well be found from traffic examination that a particular outlet receives 50 per cent of its busy-hour messages from one source whereas the remainder may be contributed from many other sources. The reduction due to the 50 per cent from one source would be equivalent to an outlet supplied from 4 sources because 25 per cent of the delay is eliminated. If the percentage of the traffic from one source rises to 80 per cent, then 64 per cent of the delay is removed. If the remaining 20 per cent of the messages all come from a second source, it will result in only 68 per cent of the delay being removed.

TABLE 5

Comparison of Delays with Different Numbers of Message Sources

MMAXI	Message Delay in Minutes								
	20 Sources			10 Sources			5 Sources		
	Minimum	Average	Maximum	Minimum	Average	Maximum	Minimum	Average	Maximum
0	Table 1	Table 1	13·61	Table 1	Table 1	13·61	Table 1	Table 1	13·32
3	Values	Values	13·61	Values	Values	13·61	Values	Values	13·32
6	Minus	Minus	15·40	Minus	Minus	15·40	Minus	Minus	15·40
9	0·26	0·24	18·40	0·26	0·75	18·40	0·26	1·38	18·40
12			21·40			21·40			21·40

5. *Abnormal Overload*

A condition for which storage equipment may have to be designed is that in which for one or more outlets there may be consistent overloading on possibly only one day of the week. It is evident that the MMAXE value will mount hour by hour so long as the overload persists. This increase will be largely determined by the extent to which the messages offered for retransmission exceed 100-per-cent loading, but there will also be a contribution due to the irregularity of the traffic flow at the time when the traffic intensity exceeds the 100-per-cent level. A reasonable idea of this increase can be estimated from Tables 1 and 3. It is evident that the overload may be accommodated by storage until the traffic slackens or by withdrawing the messages from the store for retransmission by courier or other means.

6. *Optimum Size of Store Sections*

6.1 Use of Store Sections

Torn-tape telegraph switching systems involve the storage of messages, but the very nature of the record made on punched paper tape ensures that the tape can be used only for a single message and as a consequence the provisioning of tape becomes a question of how many rolls will be required per week or month rather than how much simultaneous storage capacity is necessary.

Electronic stores have the inherent advantage that they can be used over and over again, and in consequence there arises the question of how much capacity will be required for the peak traffic condition.

It is quite unrealistic to make an electronic store with a number of sections each of sufficient capacity to accept a message of maximum length. A reasonable compromise is obtained by determining the optimum store-section size and making arrangements for each message to use as many store sections as may prove to be necessary.

6.2 Holding Time of Store Sections

The effective holding time h of a store section for a message should include allowances for the

time needed for the identification of the addressee t_1, for the period d that elapses until the outgoing circuit is free, and for the time needed for retransmission t_2. The holding time may therefore be written as:

$$h = t_1 + d + t_2. \tag{1}$$

If the input speed is higher than the output speed, (1) remains correct, but if the input speed is lower than the output speed it may be desired to await the end of the message before allowing retransmission to commence. In this case the holding time will be:

$$h = t_3 + d + t_2 \tag{2}$$

where t_3 is the time taken to receive the message.

If on the other hand retransmission is permitted to commence at once, then the holding time will be whichever is the greater of $(t_1 + d + t_2)$ or (t_3).

6.3 Factors Influencing Store-Section Size

It is inevitable that if a message uses a number of store sections the last of these store sections will on the average be only half filled. The percentage of the storage capacity that may be wasted in this way will tend to decrease as the size of the section is reduced.

Furthermore, the time t_2 will be influenced by the store-section size, and, unless the value of d is much greater than t_2, it is likely that the total store capacity can be substantially reduced by keeping t_2 reasonably small. It will be appreciated that neither t_1, d, nor t_3 is affected by the size of the store section.

These arguments would appear to favour a very-small store section, but there are two reasons for not making the section too small. In the first place, for every store section filled it will be necessary to record where the message is continued; and in the second place it will be necessary to choose free store sections for continuation and to indicate that they are engaged until they become available for reuse. Both these functions involve an amount of hardware proportional to the number of store sections.

6.4 Comparison of Store-Section Sizes

6.4.1 *Ferrite Matrix*

It is easy to make a size comparison when apparatus such as a ferrite matrix is used for storage. Assuming that each row of the matrix contains 5 characters and that the cost of continuation and supervision is equivalent to 4 rows of storage equipment, then, using the following equation, it is possible to compare different store-section sizes with the respective "row-hour" store quantities for particular lengths of message and delay:

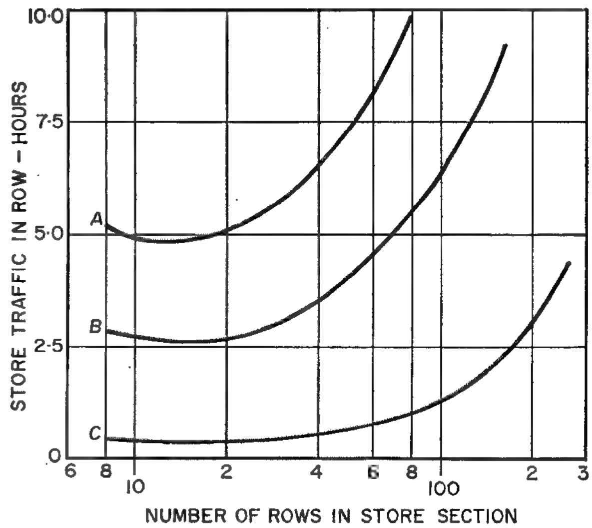

Figure 1—Determination of optimum store-section size, with a delay of 22·5 seconds between receiving and retransmitting. A = 1950, B = 1050, and C = 150 characters in each message. There are 5 characters per matrix row.

$$S = \frac{n(R+4)(d+t_1) + (R+4)(t_2)}{3600} \tag{3}$$

where

S = store traffic in row-hours
n = number of store sections taken into use
R = number of rows in store section
$d + t_1$ = delay in seconds before retransmission commences, and
t_2 = time in seconds taken to retransmit the message.

Figure 1 shows a family of curves for messages with average lengths of 1950, 1050, and 150 char-

acters, all with a delay of 22.5 seconds between receiving and retransmitting. As an example, if each store section includes 40 rows, each of 5 characters, then 200 characters can be accommodated, and for a message of 1050 characters 6 sections are needed. If each section has capacity for only 100 characters, then 11 sections are needed. However, if each section has capacity for 150 characters it is not sufficient to provide 7 sections, but 8, because it must be assumed that on the average the last section is only half filled.

The curves in Figure 1 show, for varying store-section sizes, the product of the amount of storage and the duration for which it is occupied for the 3 particular message lengths, and hence it is possible to determine the smallest load on the store for these message lengths. The optimum section size indicated by the minima of the curves in Figure 1 is expressed in numbers of rows, each containing 5 characters. The product of these two values gives the section capacity in characters and will apply also to cases in which the row is designed for more (or fewer) characters. It is evident that the longer messages need more capacity. In all cases a section of approximately 16 rows is economic and a larger section can become very wasteful.

Figure 2 shows corresponding curves for store sections with a negligible delay between receiving and retransmitting, and in this case the optimum section size decreases to about 5 rows.

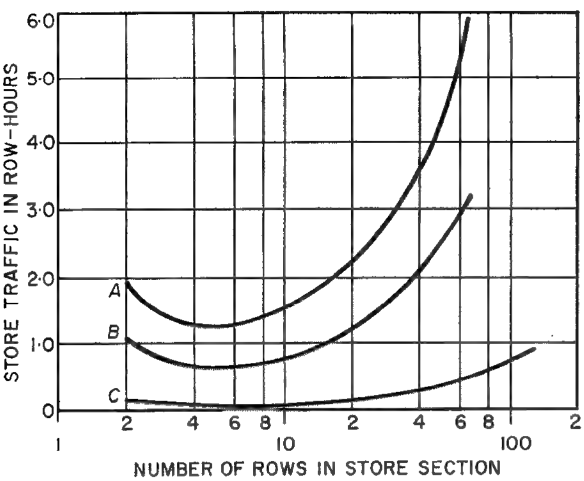

Figure 2—Determination of optimum store-section size with a delay of 2·25 seconds between receiving and retransmitting. A = 1950, B = 1050, and C = 150 characters in each message. There are 5 characters per matrix row.

Figure 3 shows corresponding curves for store sections in which the delay between receiving and retransmitting is 225 seconds; this increase tends to swamp t_2, and in consequence the optimum section size increases to about 32 rows.

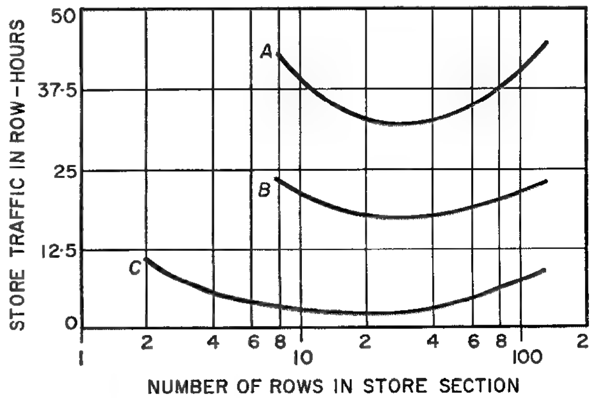

Figure 3—Determination of optimum store-section size, with a delay of 225 seconds between receiving and retransmitting. A = 1950, B = 1050, and C = 150 characters in each message. There are 5 characters per matrix row.

Figure 4 shows the optimum size of the store sections for a message of 1050 characters with varying delay between receiving and retransmitting. For the other message lengths the optimum number of rows does not vary by more than ±6.

Figure 5 illustrates 4 store sections being taken into use successively during a message. After the delay period $(d + t_1)$ the message retransmission period (t_2) commences and the store sections are released in turn. It will be seen that store section 1 could be reused for the third or fourth parts of the message. It will also be seen that the last section is used for a shorter time than the others because the message did not fill this section.

6.4.2 *Other Forms of Store*

It is a feature of a ferrite matrix that it is possible to assume a cost per row and subsequently to make a cost comparison on the basis of the row-hour traffic that is needed as a consequence of store sections of different size.

For other forms of store, such as a magnetic drum, the division of a track into a number of different sections introduces supplementary expenses that are more difficult to assess. These extra expenses will all tend to favour rather

larger store-section sizes than those shown for the ferrite matrix.

Both in the case of a magnetic drum and of magnetic tapes there are timing matters also to be taken into consideration. As an example, a

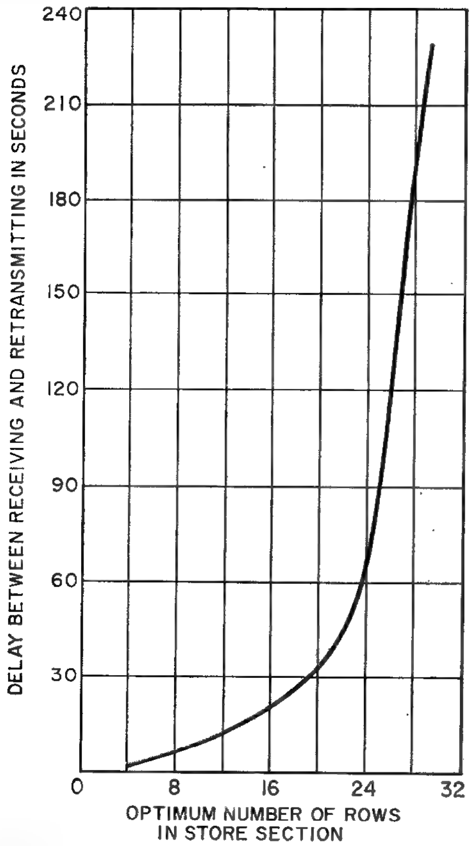

Figure 4—Optimum store-section size, with varying delay between receiving and retransmitting (1050 characters in message).

message may need to be assembled in a buffer before being transferred to a magnetic-drum store; as the store section becomes smaller and smaller, the number of transfers will increase, and it is unlikely that the duration of the transfer operation can be reduced sufficiently to prevent the larger number of transfers from causing an increase in the traffic that has to be handled by the link.

7. *Size of Common Store*

It will be appreciated that a considerable amount of data is needed to make an accurate estimate of the busy-hour traffic. The optimum size of store section does not require a precise measurement of the message length, but this length has an important influence on the average message delay. Adequate information can usually be obtained by measuring tape lengths for a number of messages. Message analysis will also provide the necessary information on the average number of characters that have to be received before the addressee can be recognised. The same analysis may also be employed to divide the average message length into the invariable pro-forma portion and the variable text portion, and also to provide an estimate of the relative number of messages addressed to one, two, three, or more addressees. This information is not likely to vary appreciably during the course of the day, but it is evident that a close scrutiny of the traffic must be made to ascertain the busy-hour period during which the store will be expected to hold most messages. Such an examination involves an hour-by-hour summary of the messages awaiting retransmission on each outlet group, and the summary will need to be taken on a succession of busy days.

After the estimate of the present-day traffic has been made it will be necessary to decide what traffic growth ought to be catered for and whether changes such as alterations in line speed

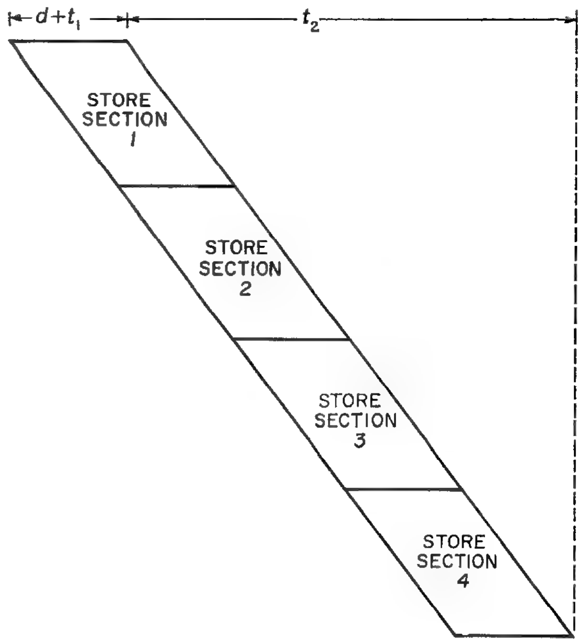

Figure 5—Relative engagement and release times of store sections during a message.

need to be taken into account. Over and above these allowances it will be necessary to make some reserve for day-to-day cases of congestion that may occur as a consequence of a failure in the line or terminal equipment.

An accurate traffic summary is difficult to produce. No two days' records are likely to be similar, and, although averages can be calculated, there is also a necessity to estimate the heaviest load likely to be experienced. This information is needed because it would be uneconomical either to cater for very-exceptional overloads or to admit frequent cases of store congestion. It is too complicated to consider giving an engaged tone when no store sections are available—a condition that may be experienced perhaps half way through a message—and it is much more convenient to make arrangements to provide an alarm when a certain degree of congestion has accumulated. To obtain relief, this alarm may be used either to restrict the incoming traffic or to initiate transfers from the common store to overflow storage machines. Such overflow machines may have a large capacity and employ magnetic or paper tape, and it is evident that the relief can be more quickly obtained if the transfer speed from the common store to the overflow store is higher than the normal line speed.

It has been explained that the common-store traffic can be calculated by (3) in terms of store-section-hours. These traffic units correspond to the call-hours in a telephone exchange and are, in effect, erlangs. By the use of well-known formulas the traffic load and a predetermined loss value will enable the quantity of store sections to be determined. In the telephone exchange example each call is considered to occupy a selector for the average holding time, whereas in the telegraph case each message is likely to occupy several store sections for a duration that may be greater or less than the message length, depending on the average delay before retransmission.

With telephone exchange traffic, message commencements and terminations are substantially unrelated occurrences, which provide something approaching a random distribution of traffic load. The use of a series of store sections for a single message introduces some interrelation, but it is not evident that the probability of instantaneous peaks will be much more pronounced as a consequence.

8. Traffic Analysis in Electronic Switching Centers

It is interesting to observe that the rapid processing possible with electronic equipment enables an automatic analysis to be made of messages awaiting retransmission for any particular direction. Furthermore, arrangements can be made to record various degrees of storage congestion so that data can be collected as to the periods and extent of the higher storage loads. Such methods are likely to prove of value in the future, on account of both accuracy and economy of effort.

9. Acknowledgments

The author acknowledges the assistance of many colleagues, in particular that of Mrs. E. Loverseed and Mr. A. D. Marr.

Shared-Channel Voice-Frequency Telegraph Equipment, UT-57/1*

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TELEPRINTERS are often operated over telephone channels and a voice-frequency telegraph equipment, *UT-57/1*, has been developed for such use. Transmission may be over wire and cable lines as well as by radio.

Transistors are used exclusively. The parts are mounted on both sides of a relay panel and include the voice-frequency telegraph equipment, power supply, and a converter for direct connection to the teleprinter. The channel equipment uses 6 transistors. Level variations of ±0.7 neper (±6 decibels) are compensated by a fast-acting level corrector. The measured distortion during operation is ⩽5 percent and, hence, is in accordance with the recommendations of the Comité Consultatif International Télégraphique et Téléphonique for standard telegraph channels.

1. *Mode of Operation*

1.1 Combining of Teleprinter and Telephone Channels

The frequency of 2820 cycles per second, which is channel *21* of a 24-channel voice-frequency telegraph system, was selected for the carrier because some telephone channels, for instance those in radio equipment, often show a substantial increase in attenuation at 3000 cycles. To provide for the telegraph band, the high-frequency limit of the speech band was reduced to 2400 cycles. The resulting loss of syllable articulation is negligible and hardly affects the understandability of speech transmitted over the system.

The frequency scheme employed is shown in Figure 1. At both the transmitting and receiving ends, band-elimination and band-pass filters are used rather than the conventional low-pass and high-pass filters. The band-elimination filter for the suppression of speech frequencies is simpler than a comparable low-pass network and the band-pass filter can be utilized as a transmitting and receiving filter for the telegraph channel. Unlike a low-pass filter, however, the band-elimination filter does not suppress frequencies above 3300 cycles. A highly resistive attenuator decouples the band-pass from the band-elimination filter. The telegraph transmitting level must be adjusted with respect to the speech level to keep interference between the two channels as low as possible. The direct leakage from one system into the other must be kept small by providing adequate attenuation in both the band-elimination and band-pass filters. If this attenuation is insufficient, speech frequencies appearing in the telegraph channel cause irregular telegraph distortion. On the other hand, the keying spectrum generated by the modulation of the telegraph carrier will reduce the signal-to-noise ratio in the telephone channel if the band-pass filter does not properly attenuate the frequencies outside of its pass band. Both effects can be minimized by suitable dimensioning of the filters.

The disturbances caused by nonlinearity of the transmission system are of a more-serious nature. The third harmonics of all speech-frequency components lying near 940 cycles generate interference in the telegraph channel. This is true also for the second harmonics of frequencies around 1410 cycles. Harmonics generated in the common transmission path are not affected by the filters and the only remedy is to make the telegraph transmitting level high compared with the speech level, for a given distortion coefficient of the transmission system. Nonlinearity of the transmission system will also reduce the clarity of speech through the generation of spurious frequencies resulting from the beating of telegraph frequencies with various voice frequencies during simultaneous transmissions. Also, a high telegraph level re-

* Originally published in German under the title, "Das Einkanal-Überlagerungstelegraphiegerät UT-57/1," in *SEG-Nachrichten;* volume 6, number 3, pages 146–149; 1958.

duces the damaging effects of line noise on the telegraph distortion. On the other hand, to assure a good signal-to-noise ratio in the speech path, the telegraph level must not predominate. The optimum telegraph level was found to be 0.2 neper (1.7 decibels) below the reference level of

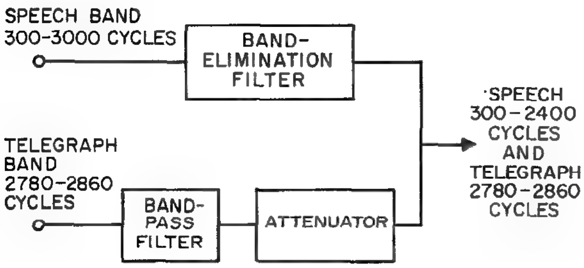


Figure 1—Arrangement of telephone and telegraph frequency bands in a telephone channel.

the speech path. Since this zero level is rarely reached in normal speech, the telegraph level is relatively high. These signal levels permit the use of transmission paths subjected to substantial distortion and strong interference as might be encountered in radio links. The band-pass filter was so dimensioned that a signal-to-noise ratio of 6 nepers (52 decibels) is safely maintained despite the relatively high telegraph level.

The sum of the telegraph and speech levels must not exceed the point of overload of the transmission system; otherwise, inadmissibly high distortion would result. To eliminate this danger, a limiter was provided at the input of the band-elimination filter to prevent any speech peak from exceeding the reference zero level by more than 0.2 neper (1.7 decibels).

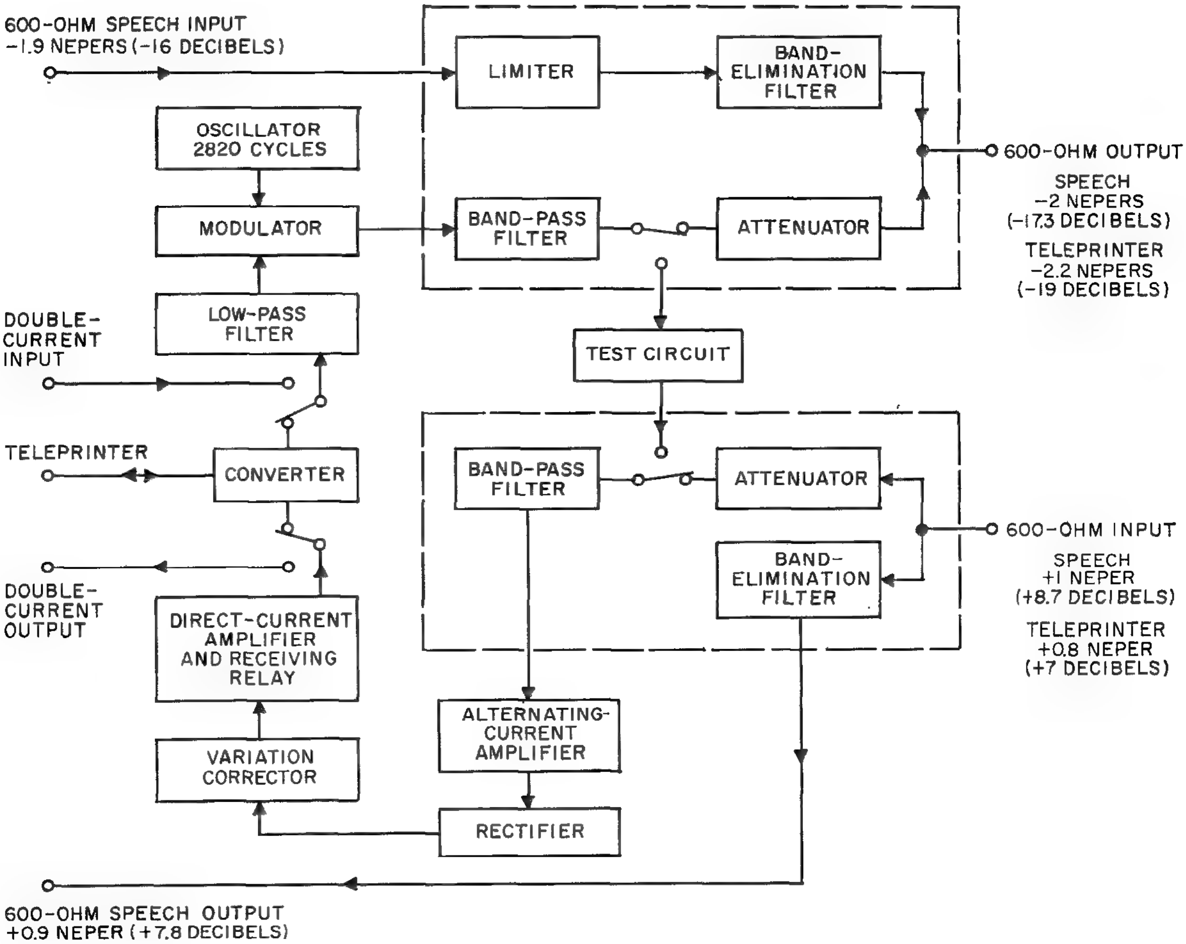


Figure 2—Block diagram of *UT-57/1*.

1.2 Transmitting Equipment

A block diagram of the telephone and telegraph combining equipment, *UT-57/1*, is shown in Figure 2. An oscillator employing a junction transistor in a bridge circuit generates the carrier. The frequency-determining resonant circuit is temperature compensated. The oscillator output voltage is stabilized so as to be independent of power-supply fluctuations. Its output can be adjusted over a range of ± 0.2 neper (± 1.7 decibels). The oscillator is connected to the modulator for keying the amplitude of the carrier.

The double-current or polar signals coming from the converter pass through a low-pass filter that removes the short interruptions caused by contact chatter of the transmitter relay or by the armature reversal. This ensures a steady transition between spacing and marking currents. The output signals thus formed control a push-pull combination of four diodes in the modulator so that the modulator conducts for one polarity of voltage and attenuates by at least 6 nepers (52 decibels) for the other polarity. This results in full carrier suppression. Another two diodes are responsible for the so-called interruption attenuation by means of which the carrier is completely blocked if one or both conductors of the local circuit are opened. The modulator is wired for open-circuit operation; that is, a tone goes to the line only when actual writing takes place. However, the equipment is easily converted for closed-circuit operation.

The intrinsic distortion of the modulator is low. If the current in the local circuit is reduced from its rated value of ± 20 milliamperes to ± 10 milliamperes, the resulting additional telegraph distortion is negligible. The diodes are so arranged in the modulator that its output impedance has the same value both in the conducting and in the nonconducting state. As a result, the band-pass filter connected to the output is terminated with the correct source impedance at all times. This is a triple-tuned filter that has a bandwidth of about 75 cycles between the half-power points and is dimensioned so that the signal-to-noise ratio in the speech channel complies with the requirements. The signal spectrum appears at the output of the equipment after having passed through an attenuator serving as a decoupler.

1.3 Receiving Equipment

In the receiving branch, the telegraph signal passes through a decoupling attenuator and a band-pass filter, which is an exact duplicate of the band-pass filter in the transmitting branch. It then goes to a three-stage transistor alternating-voltage amplifier. The amplifier input contains a continuously adjustable level control by which the receiver sensitivity is adjusted to the center of the range over which the automatic level corrector operates. A jack permits this center level to be checked with a voltmeter. A push-button switch in the negative-feedback branch decreases the level by 0.5 neper (4 decibels) to facilitate adjustment of the polarized receiving relay to its neutral position.

The amplified telegraph signal is then demodulated in a full-wave rectifier and applied to a level-correcting unit. The originally square amplitude-modulated telegraph signal is distorted into pulses with almost sinusoidal leading edges as a result of the frequency limitation imposed by the two band-pass filters. The proper pulse duration remains at only half the pulse amplitude. To prevent telegraph signal distortion, the armature of the polarized receiving relay must reverse just when the signal pulse has reached half its amplitude after a change in polarity. However, the signal amplitude will vary somewhat, which requires that this feature of the operation of the relay be under control of the received-signal level. Conventional control circuits employ components that produce integration and they cannot therefore compensate instantaneously for sudden level fluctuations. For this reason, an instantaneously responding level corrector was provided. A network derives a correcting voltage with the required amplitude by differentiating the received signal voltage. This voltage is applied via a direct-current amplifier to a separate opposing winding of the receiving relay and immediately corrects the level for each individual signal-pulse edge. By alternately operating two controls, the corrector circuit can be adjusted so that the telegraph distortion practically disappears over a level range of ± 0.7 neper (6 decibels).

The correcting unit is followed by a direct-voltage amplifier having two transistors. This difference amplifier has two inputs one for the signal and one for the correction voltage. Its output is connected to the receiving relay.

1.4 Power Supply

The equipment consumes 25 watts and can be operated by batteries providing +60 volts and −60 volts to ground. For convenience, the built-in power supply develops these voltages from the alternating-current power line of either 110 or 220 volts with a tolerance of ±10 percent and at frequencies between 48 and 65 cycles.

1.5 Operating Features

The quality of a telegraph transmission system is judged primarily by the magnitude of telegraph signal distortion. Before measuring, the channel should be adjusted to neutral. For this purpose, a switchable test path has been provided connecting the transmitting band-pass filter via an attenuator to the receiving band-pass filter input; an alternating-polarity test voltage having equal amplitude and duration for each polarity is applied to the near-end input of the modulator, and the double-current pulses appearing at the output of the receiving relay go to an instrument or distortion-measuring set to produce an optical display. Then the controls of the level corrector are so adjusted for levels of 0 and −0.5 neper (0 or −4.3 decibels), respectively, that the receiver is in its neutral position.

After system adjustment to the neutral state, tests of the overall distortion with the Comité Consultatif International Télégraphique et Téléphonique test text at 50 bauds showed values of less than 3 percent in the level range ±0.7 neper (±6 decibels). This distortion did not exceed 5 percent even when the operating voltage fluctuated by ±10 percent and the carrier frequency was detuned by ±7 cycles. The receiver responds to a level of about −1.55 nepers (−13.5 decibels) referred to the level range center. Interfering voltage having peaks of less than this value cannot cause faulty operation even in the keying intervals.

1.6 Teleprinter Connections

To facilitate directly connecting the teleprinter to the equipment, a simple converter from two-wire single-current teleprinter operation to four-wire double-current equipment and line operation has been provided. The circuit is shown in Figure 3. A battery provides a steady-state current of 40 milliamperes to the teleprinter.

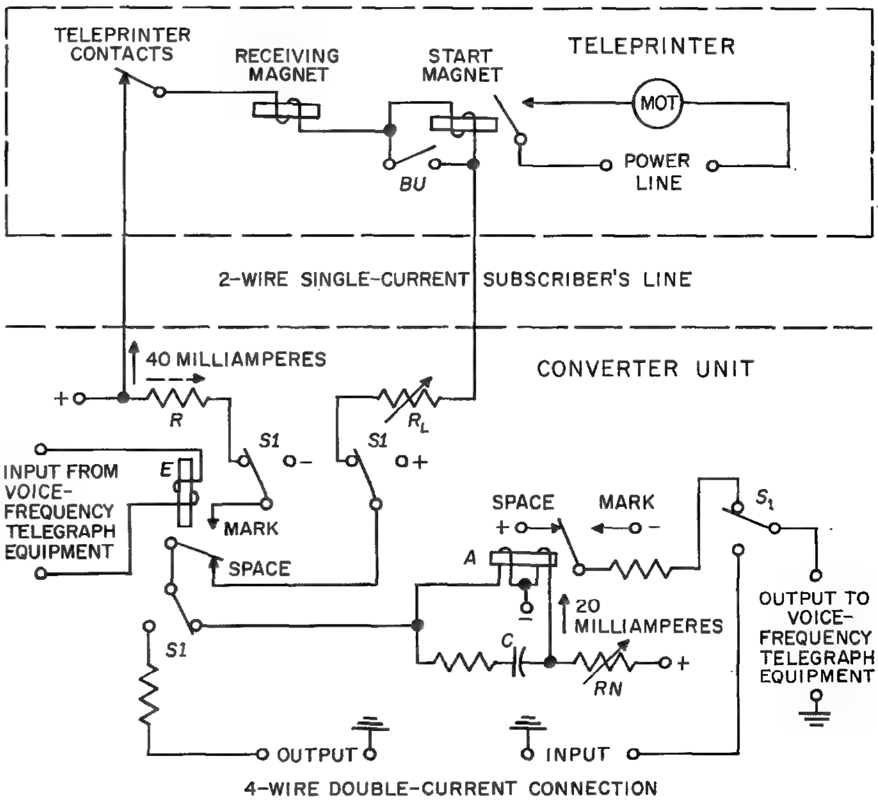


Figure 3—Converter permitting teleprinter operation from either a two-wire single-current or a four-wire double-current system.

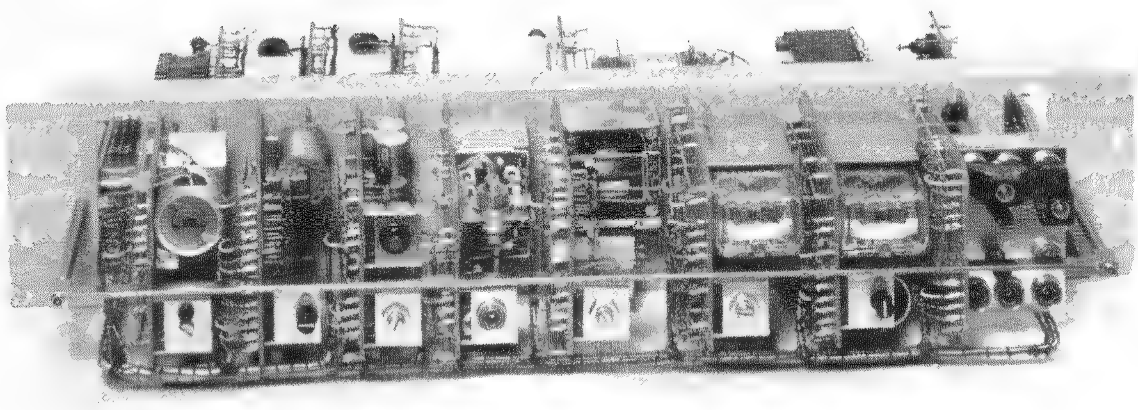


Figure 4—Front view of the *UT-57/1* equipment with covers removed.

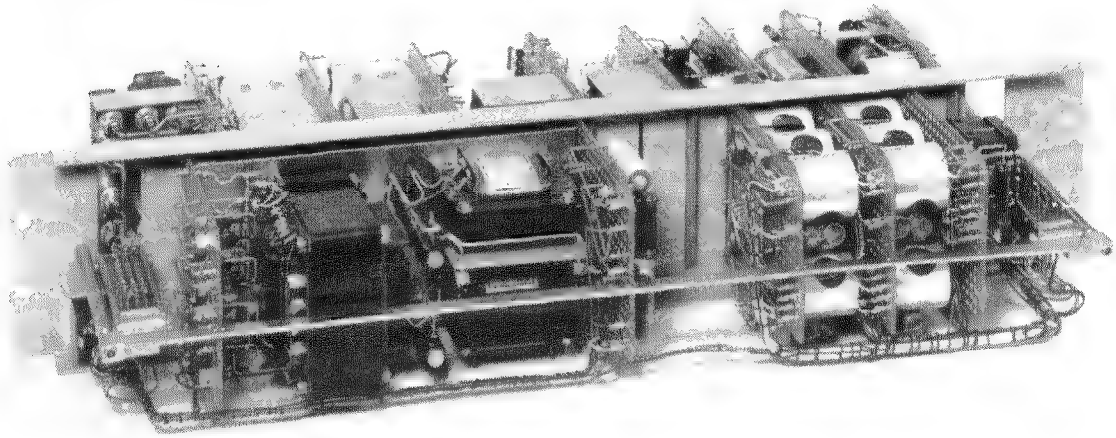


Figure 5—Rear view with covers removed.

To obtain this rated current, the effects of the line losses can be compensated for by R_L.

By depressing the teleprinter key BU, the solenoid of the motor switch is short-circuited and its contacts close and start the motor. During typing, the teleprinter transmitting contacts open and close the subscriber loop according to the code of the telegraph signals. When the subscriber loop is closed, the flow of current through the left-hand winding of relay A in the converter actuates contacts that put $(+)$ on the modulator of the voice-frequency telegraph equipment through the low-pass filter. In the other winding of this telegraph relay, an opposing current of 20 milliamperes flows so that when the subscriber loop is open and no current flows in the first winding, the relay operates to the other side to send $(-)$ to the modulator, which releases the 2820-cycle signal to the outgoing line.

At the far-end station, the corresponding receiving relay E is reversed in polarity and its contact is switched to the marking position. The relay A in that station is not affected because, in marking position, a current of 40 milliamperes flows through resistor R and the moment of reversal is filled by the energy stored in capacitor C. Through the reversal of the contacts of E, the current through the teleprinter receiving magnet is interrupted in the rhythm of the telegraph signals. The first current interruption causes the teleprinter motor to be started. This motor continues to operate until a writing interval exceed-

ing 30 seconds occurs. The resistor *RN* serves to adjust the opposing current in relay *A* and, hence, establish the neutral position of the latter. Switch *S1* changes the circuit to permit either two- or four-wire operation.

An additional toll subscriber relay terminal strip is provided for operation through manual or automatic exchanges. It is designed so that, apart from the actual message, those signals necessary for establishing a connection in toll traffic are converted into a form suitable for voice-frequency telegraphy. The subscriber station can be connected for either four-wire single-current duplex or two-wire single-current simplex operation. The operating voltages required for this additional strip are derived from the power supply incorporated in the *UT-57/1* equipment.

2. Constructional Details

System constructional design was aimed particularly at simplicity. On both sides of a conventional relay panel 520 millimeters (20.5 inches) wide and 100 millimeters (4 inches) high, pertinax boards are fixed to which the individual units are mounted. The boards are first completely wired and then fastened into slots in the panel. They are then interconnected by preformed cable harnesses whose terminals are soldered to the boards. This construction combines simplicity of assembly with easy accessibility and facilitates replacement of parts. Individual boards can be exchanged for others meeting special requirements. The wiring can be altered readily since all boards have a standard raster of holes providing for the simple insertion of soldering lugs in any desired arrangement. The weight of the completely equipped *UT-57/1* panel is about 10 kilograms (22 pounds).

Figure 4 is the front view of the panel. From left to right, are the oscillator with transmitting level control, modulator and test path, input stages of the alternating-voltage amplifier with control and 0.5-neper push-button key, power stage of the alternating-voltage amplifier with toll jack, rectifier and level corrector with the second neutral control, direct-voltage amplifier with receiving relay *E* and the first neutral control, converter with relay *A* and reversing switch, and jacks for the subscriber station. The rear side of the panel as shown in Figure 5 mounts the power supply, the transmitting and receiving directional filters, and the limiter. The components are protected by removable covers,

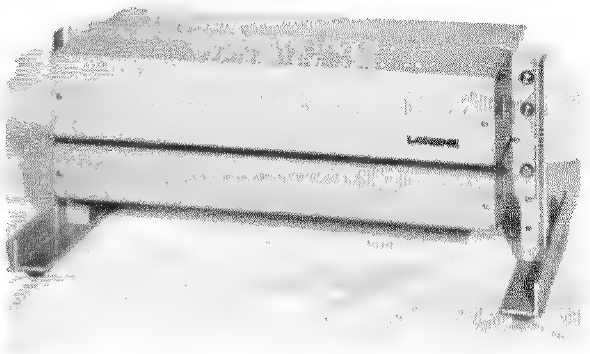


Figure 6—The main equipment is housed in the upper covers and the exchange-operation strip is below it.

which will be seen in Figure 6. Below the panel in this illustration will be seen the relay terminal strip required for exchange operation.

3. Compatibility with Radio Equipment

The voice-frequency telegraph equipment was tested with an *SEF-7* very-high-frequency radio equipment[1] built by Lorenz. The setup is shown in Figure 7. This radio equipment employs frequency modulation. To prevent interference to adjacent radio services, a limiter prevents excessive frequency shift. At the design level the shift is ±10.5 kilocycles. Through the action of the limiter, the maximum swing is ±15 kilocycles at +0.4 neper (+3.5 decibels). The telegraphy signal in the speech channel causes an increased level to appear at the input of the radio-frequency modulator; therefore the frequency shift for the reference level must be reduced to about 7 kilocycles by a series input attenuator of about 0.4 neper (3.5 decibels). This gives a sufficient margin between the summed message (speech and telegraphy) and the overload level. Measurements have shown that this reduced swing does not essentially reduce the range of the radio equipment. A signal-to-noise ratio of about 3 nepers (26 decibels) in the speech path at the

[1] G. Sidow, "UKW Funkgerate für bewegliche Funkdienste," *SEG Nachrichten,* volume 5, number 2, pages 79–87; 1957.

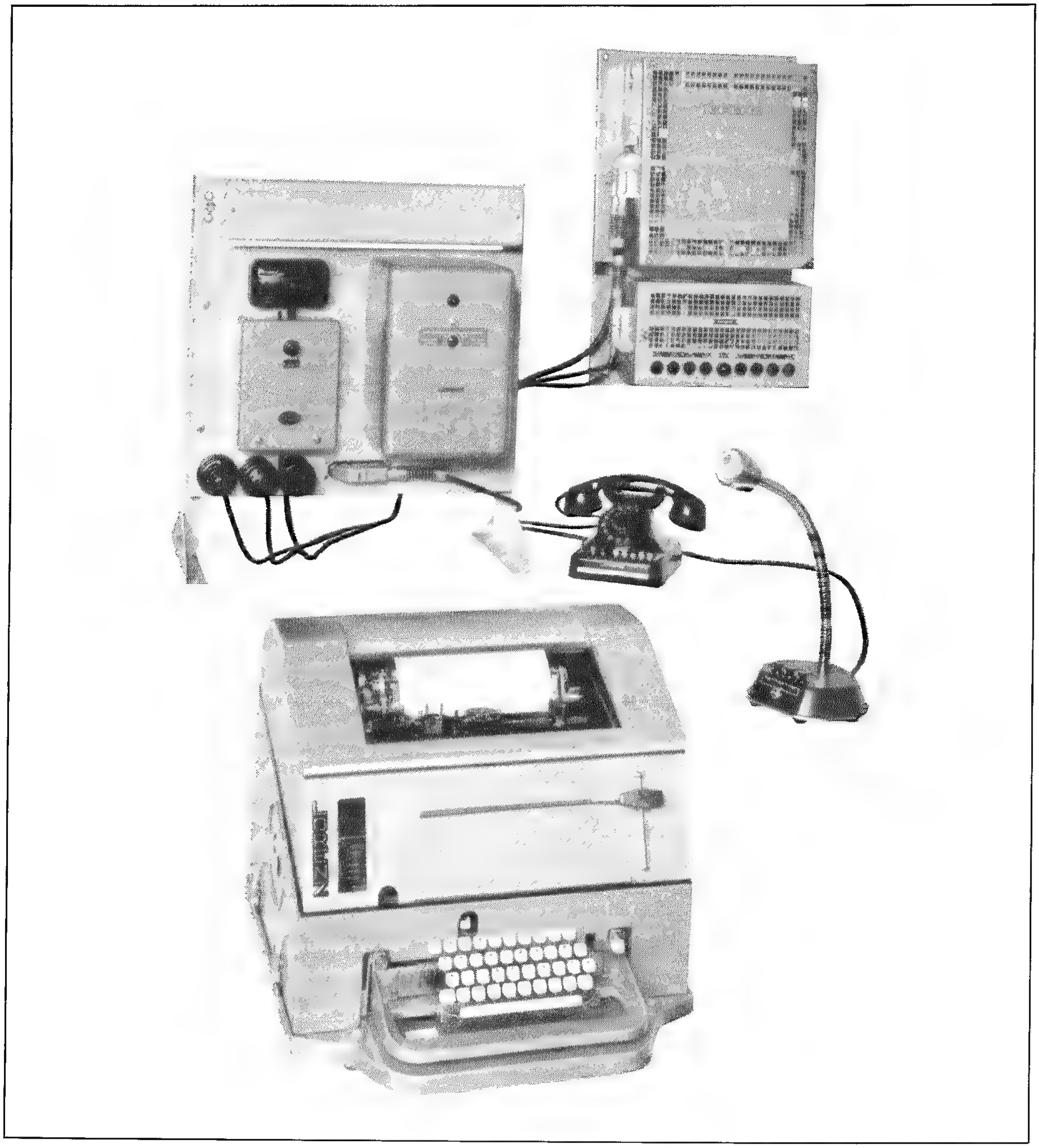


Figure 7—Test setup with the *SEF-7* very-high-frequency radio equipment. The *UT-57/1* strip is mounted on the transfer-equipment frame just back of the teleprinter.

receiving branch is still sufficient for practically undisturbed teleprinter operation. In the *SEF-7* receiver, this corresponds to a receiving field strength of about 1 microvolt per meter. White noise was used as interference in these measurements. The noise suppressor built into the radio equipment must not be switched off in voice-frequency telegraphy operation because absence of the transmitting carrier will increase the sensitivity of the radio receiver and the interference noise present will be powerful enough to start the teleprinter.

Development of Creed Telegraph Apparatus

By BRUCE BROOKE-WAVELL

Creed & Company, Limited; Croydon, England

MOLIÈRE'S character, M. Jourdain, awoke one day to the realization that he had been speaking prose all his life without knowing it. Similarly, telegraph engineers have recently discovered, thanks to Claude Shannon and others, that they have been processing digital information for the last hundred years without being aware of that fact. Fortunately for telegraph manufacturers, many other people have discovered this fact too, even if in some cases they have required a little prompting.

The result has been, during the past five years, that a wide range of telegraph apparatus that was originally developed for purely conventional telegraph applications has been used for quite different purposes. The punched-tape technique, for example, that was originally introduced to save line time is now employed for providing input and output facilities for digital computers, for data recording and processing, and for the automatic control of machine tools.

Back in 1948, however, at the beginning of the ten-year period being reviewed, telegraph engineers were too fully occupied with the pressing post-war requirements of conventional telegraphy to give much attention to other matters. This was certainly true at Creed where, during the war, production was for strategic reasons concentrated on a small range of telegraph machines and in particular on the model-*7* page teleprinter, which accounted for over half of the total number of machines turned out. There was, therefore, the urgent problem of pushing ahead with a variety of new developments to meet the needs of both British and overseas telegraph administrations.

1. Early Post-War Development

The first such development was for the British Post Office which started to reorganize its public telegraph service. To replace the pre-war model-*3* tape teleprinter, which since its introduction in 1929 had been in general use for the transmission and reception of telegrams, the model-*47* tape teleprinter was produced. This machine was similar in principle of operation to the model *7* page teleprinter but embodied improvements resulting from experience with that model and from Post Office experience with the earlier tape machine. Perhaps the most important difference between the models *47* and *7* teleprinters, apart from the obvious one that the page unit on the latter machine had been replaced by a tape unit on the former, was that the model *47* had a 4-row, typewriter-pattern-layout, saw-tooth-comb bar keyboard in place of the earlier 3-row motorized keyboard.

There next followed a series of developments concerned with the provision of automatic tape transmission facilities for the full utilization of line time. This was to meet the increasing interest being shown in simple tape preparation and automatic transmission sets for use both in individual installations and in the more-ambitious manual-transfer push-button semi-automatic tape relay systems.

Equipment for automatic tape transmission sets had been in production since before the war, including a keyboard perforator, a nonprinting reperforator, and an automatic tape transmitter, but a further range of machines was now introduced mainly for use in tape relay systems. The model-*85* printing reperforator was developed for recording messages on printed chadless tape, that is, perforated tape with the chads remaining attached and with the printing superimposed on them. The printing was provided to eliminate the need for operators to read the 5-unit code and the chadless method of perforating to retain the use of standard-width tape. A variant of this machine—the model-*86* printing reperforator—recorded messages on fully perforated wide tape with the printing underneath the perforations.

An essential requirement in tape relay stations is the use of ganged automatic tape transmitters. To meet this requirement, models *71*, *72*, and *74* multiple-tape transmitters were produced. The standard single-head transmitter, model *6S*, was

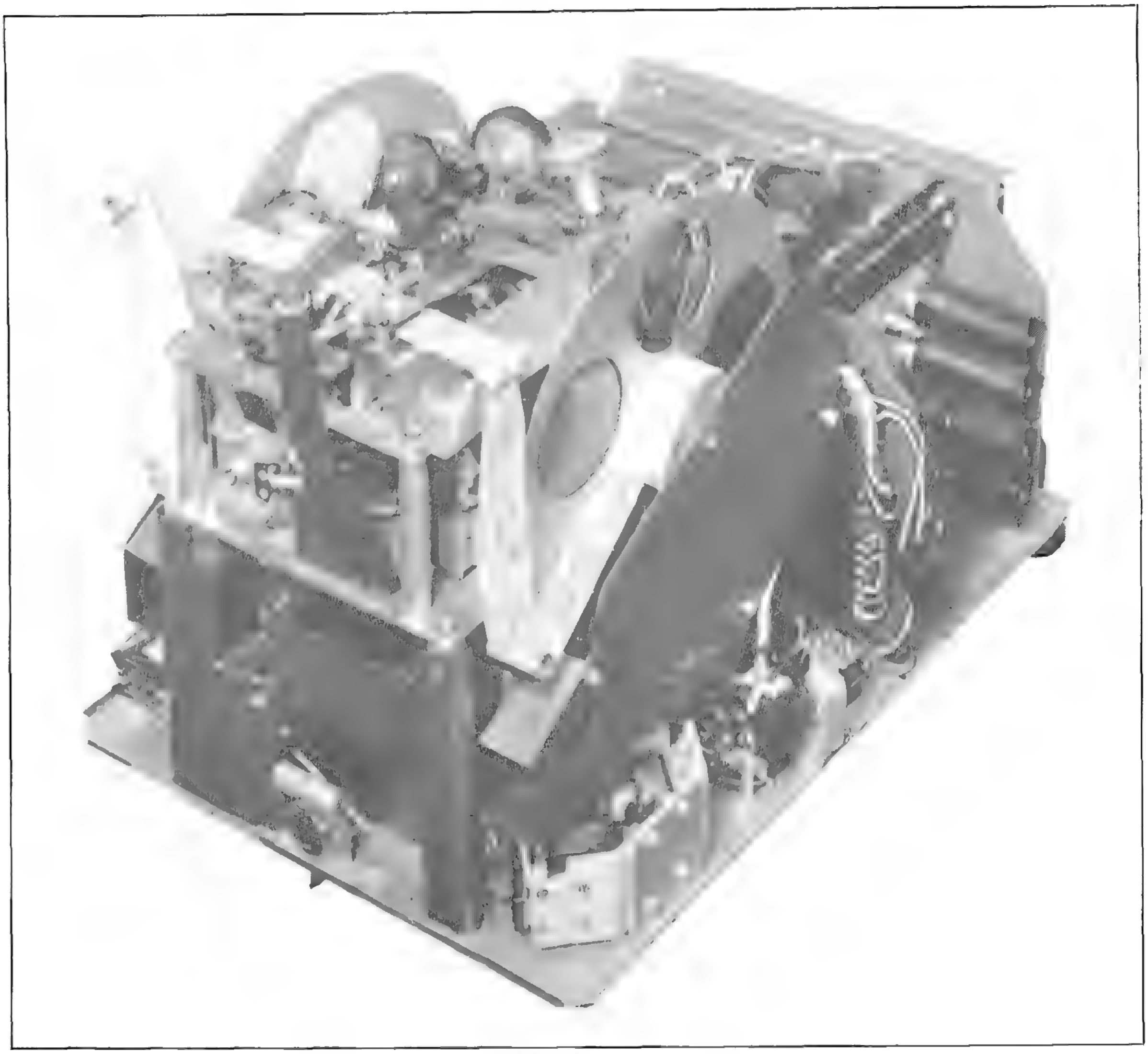


Figure 1—A special variation of the model-25 reperforator. It provides a punched-tape record of the routine operations of equipment such as cash registers and accounting machines for computer input.

redesigned, a transmitting mechanism of the striker pattern being fitted to it to ensure high-quality signals.

Meanwhile, the increasing use of teleprinters by business concerns renting circuits from the British Post Office, as well as the changing needs of telegraph administrations, made it necessary to add to the model-7 page teleprinter, which was still in great demand, a variety of new facilities.

This machine had been introduced in 1931 as a simple, sturdy teleprinter in which the design emphasis had been on simplicity of manufacture and maintenance rather than on a large number of facilities. These characteristics were, undoubtedly, of great value to the armed services and the British Defence Teleprinter Network in which the bulk of the war-time production was used. The changed circumstances after the war, however, created new needs.

The machine was designed originally for double-current (polar) operation: it was now modified to make it equally suitable for single-current (neutral) operation. A new cam unit was fitted to it having an orientation device, that is, a means for centralizing the margin and

for measuring the receiver tolerance in the absence of a separate margin-testing set. Later, a further alternative cam unit was made available to provide for immediate printing and to increase the receiver tolerance to over 86 per cent. A reperforating attachment was fitted that could be used to prepare perforated tape with a local record or provide a perforated record of incoming messages as required. Two-colour printing, a period-of-operation counter, improved visibility of the printing point, better sound-proofing arrangements, and many other changes were introduced, transforming the teleprinter into what superficially appeared to be quite a different machine. As the original simple teleprinter continued to sell to customers who did not need these new facilities, the modified machine was distinguished as being the new model *54*.

2. *New Applications*

About five years ago, as already mentioned, a series of entirely new markets with requirements quite unknown to the telegraph art began to appear. Digital-computer manufacturers turned to punched tape as a way of solving their input and output problems. The development of the techniques of data recording and processing, as well as of automation, created a demand for further kinds of punched-tape equipment.

To meet the need of the computer manufacturers, a comprehensive range of *tape-editing* equipment in the form of selfcontained comparator, verifier, and reproducer sets were developed. These consisted of standard teleprinter equipments to which new facilities had been added, together with specially designed relay units and control panels. A series of interpreter sets was also produced to provide computer print-out facilities.

It was early realised that while normal telegraph speeds would be more-or-less adequate for tape-editing equipment, much higher speeds were required for the input and output devices. The computer manufacturers found a satisfactory method of reading input tapes by using photoelectric means for registering the holes. Creed did not, therefore, develop a high-speed tape reader.

The output problem, however, was a more-difficult one that was approached initially by introducing the model-*7P/5W* reperforator. This was derived by modifying a keyboard perforator to reperforate from a 5-wire input. It operated at 15 characters per second, thereby permitting recording at slightly over twice the speed formerly possible with standard teleprinter equipment.

The next step was the introduction of a tape punch that was specifically designed for output recording. This model-*25* reperforator records data in 5-, 6-, or 7-track tapes—either in single tapes or in two tapes at a time—at a speed of 33 characters per second, which is 5 times the normal teleprinter speed. The model-*25* reperforator is at present the standard output punch for the majority of British electronic digital computers. Over 1000 of them have been produced to date for this and other applications, which include the automatic logging of telephone calls on punched tape and the provision of a punched-tape record of the automatic routine operations of a range of modern business equipment such as cash registers and accounting machines. Figure 1 shows the mechanism of one of these models.

The speed of the model-*25* reperforator is still not high enough, however, to make it a completely adequate computer output recorder. Two further machines have, therefore, been developed for this purpose. The first, the model-*3000* reperforator, is the world's fastest output punch, operating at a speed of 300 characters per second. It may be seen in Figure 2. The other is a hydraulically operated character-by-character printer with a speed of 100 characters per second. This operates from a 5-wire input, the characters being built up on the mosaic principle with a 5 by 5 grid. As it has 3 times the speed of the model-*25* reperforator, it is advantageous to use it as a direct-output printer. Alternatively, it may be used in conjunction with the model-*3000* reperforator if a higher output speed is required. The speeds obtainable with these machines are adequate for most purposes. Since they are much-less expensive than the higher-speed line printers on the market, they will undoubtedly satisfy an important need in the computer field. Figure 3 is an example of the use of teleprinter equipment adapted to computer requirements.

A range of machines is also under development in the data processing and automation fields. These are, in most cases but not all, modified versions of standard teleprinter equipment and

include a machine for recording 5-unit combinations on the edges of cards and tickets, a tape transmitter for process control by punched tape, which has been modified to feed forwards or backwards under the control of external signals, and a teleprinter that operates directly from a 5-wire input.

Figure 2—Model-*3000* tape punch. It was designed for use as the output equipment for electronic computers. It can punch 5-, 6-, 7-, or 8-track tape at the high speed of 300 characters per second or 3000 words per minute.

All these are individual instruments that have been produced for use in conjunction with other manufacturers' equipment. Creed is also investigating the problem of providing complete integrated data-processing systems using punched tape throughout, for use where the total quantity of information to be processed is not sufficient to justify the employment of electronic equipment.

The next step will almost certainly be the extension of both the computer and data-processing fields to include telegraph communication facilities. When this occurs, the demand for higher speeds of operation, codes with a greater information-carrying capacity, and new standards of transmission accuracy will radically affect present telegraph practice.

3. Fresh Start in Teleprinter Design

The history of the models *7* and *54* teleprinters and the rapid growth of nontelegraphic applications of telegraph apparatus both played an important part in determining the design of the new model-Seventy-five teleprinter.

Some four years ago, it was realised that the model-*54* teleprinter, while still adequate for normal communication needs, could not be modified to meet the new requirements that were arising in the communication and data processing fields without undue complication and increase of weight.

To take the communication field first, it was becoming evident that with the accelerating expansion of the telex and private-wire services, the teleprinter was becoming an indispensable piece of office furniture on the same basis as, although complementary to, the telephone. There was reason to believe that the large telegraph organisations, both public and private, would tend in the future to provide the interconnection facilities such as switching and channel equipment, but leave the operating and servicing responsibilities to the user.

From this the conclusion was drawn that teleprinters would have to meet, in the near future, the following requirements:–

A. *Reduced Maintenance.* The dispersal of telegraph equipment in business offices instead of their previous concentration in telegraph offices would put a premium on low maintenance.

B. *Dual-Speed Operation.* For economic reasons, there would be a strong demand for automatic tape transmission sets. An improvement on present practice could be made if a dual-speed teleprinter with a manual speed-change control were used that could prepare tape "off line" at a high speed and provide message transmission facilities "on line" at the normal telegraph speed.

C. *Modern Appearance and Small Size.* If teleprinters were to become standard office furniture, their appearance and size would play a more important role than in the past.

These requirements were seen to hold also for the data-processing field, but there were additional requirements. It was apparent that an increasing demand would exist for communication channels between business offices and centrally placed computers serving them. Many of these business offices would not want full-time channels but would prefer channels that could be used partly for transmitting data and partly for normal communication traffic. There would, therefore, be an advantage in using dual-purpose terminal equipment that would be suitable for the transmission and reception of messages and also for the local processing of data. As these two kinds of use ideally require sequential and simultaneous (5-wire) modes of signalling respectively, a satisfactory dual-purpose machine would have to be readily convertible from one mode of signalling to the other.

All these requirements have been met in the design of the model-Seventy-five teleprinter, which needs less-frequent maintenance than previous Creed teleprinters, operates at both 66 and

Figure 3—Application of teleprinter equipment to record output data from an electronic digital computer in punched-tape and printed-page form.

100 words per minute, is the smallest teleprinter in the world in production, has a smoothly styled modern appearance, and is easily convertible to either 5-wire or sequential operation.

This machine is shown in Figure 4. It has been in production for some months and is the basic unit of an integrated range of telegraph machines that are now in various stages of development. A reperforating attachment for the model-Seventy-five teleprinter is already in production, whilst a new automatic tape transmitter attachment, a printing reperforator, and a tape teleprinter are on the way.

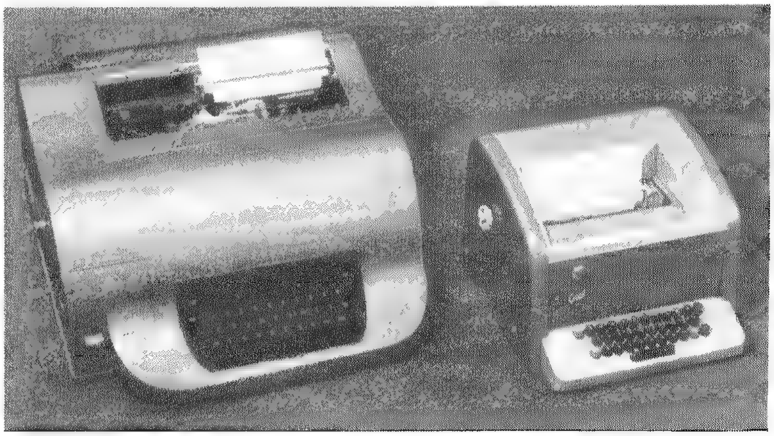


Figure 4—The model-Seventy-five teleprinter at the right is only about half the size and a third of the weight of former models. It is capable of operation at 100 words per minute with greatly extended servicing intervals.

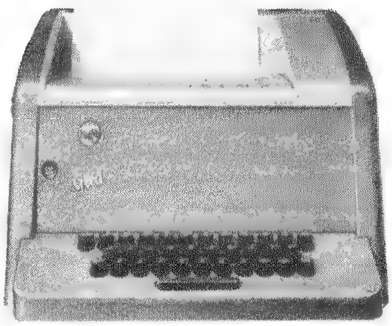
Mention of the model-Seventy-five teleprinter brings us to the present phase of Creed development. As this is, naturally, of greater interest than the largely historical material in this article, a full, separate article has been devoted to it.

Creed Model-Seventy-Five Teleprinter

By BRUCE BROOKE-WAVELL

Creed & Company, Limited; Croydon, England

EARLY IN 1958, the first production units of a new teleprinter left the Creed factory. This model Seventy-five machine (Figure 1) is the smallest teleprinter in production in the world, and its weight of 35 pounds (16 kilograms) for the receiver-only version is well under half that of most other teleprinters.



Considerable interest in these features was shown by aircraft operating companies well before the production stage was reached. In March, 1956, a pre-production version of the machine was fitted as an airborne teleprinter in a British Overseas Airways Corporation Stratocruiser on regular service from London to New York and continuously recorded weather data and other essential flight information broadcast from radio stations in Galdenoch, Scotland, and Halifax, Nova Scotia. This trial was very successful, the operation of the machine being unaffected by either vibration or tilting.

All this might suggest, on the principle that one cannot get something for nothing, that small size and light weight have been obtained at the expense of robustness, range of facilities, standards of performance, or other features. This, however, is not so. It is the first and basic unit of a new integrated range of teleprinter equipment to supersede existing machines, both in conventional telegraph communication services and in the growing number of special applications in data processing and automation. The surprising reductions in size and weight are byproducts of new design methods.

The basic design of the model-Seventy-five teleprinter is quite different from that of present machines. Existing Creed teleprinters were developed by modifying and adapting the model 7, which was originally introduced in 1931, to provide the greater variety of facilities and higher standards of performance called for by progress in telegraphy since that time. Although these machines are adequate for most present needs, it is becoming increasingly difficult to add new facilities to them without also adding disproportionately to their size and weight. By making a completely fresh start, it has been possible to provide the full range of present-day facilities in a much-simpler and more-direct manner while, at the same time, making full allowance for foreseeable future requirements.

1. General Features

The teleprinter has been designed to operate reliably at 100 words per minute to cater for an increasingly widespread requirement, especially in data processing and computer output applications. At the more-commonly used communication speeds of 66 and 60 words per minute, to which the teleprinter can be adapted by a simple gear change, this reserve speed provides a large additional safety factor.

Single- and double-current operation are both catered for, the minimum receiver operating current being 40 and 20 milliamperes, respectively. The same transmitter contact assembly, without requiring any relay, is used for both modes of operation. The 5 code elements are first set up simultaneously on 5 changeover contacts and then transmitted sequentially by 5 make–break contacts. This use of separate transmitter contacts for each code element gives optimum transmission characteristics on single-current circuits.

The basic operating facility provided is simplex plus local record, the local record being obtained by a direct mechanical connection between the keyboard and receiver and not, as on previous machines, electrically through a leak resistor. The transmitter changeover contacts read off the transmitted combinations from the selecting pins on the receiver and the make–break contacts are

Figure 1—Smaller, lighter, and simpler than any other teleprinter in production in the world, the model Seventy-five is capable of sustained operation at 100 words per minute with high standards of attention-free performance.

operated by cams on the receive cam sleeve. The need for a separate keyboard transmitter is thereby avoided. This feature is responsible for a substantial reduction in the number of parts in the teleprinter. A further advantage is that the local record is printed immediately after the depression of a key, without any perceptible time lag.

The keyboard can be of the 3- or 4-row type, and the several hundred keyboard layouts catered for on previous Creed teleprinters are all available on the new machine.

The teleprinter is constructed on the unit principle, which has been proved by experience to result in simpler maintenance and manufacture. All units are interchangeable with equivalent units on other machines provided they are adjusted.

Special care has been taken to reduce the maintenance required; the lubrication intervals, for example, being extended to approximately 1000 hours of operation, compared with 300 hours for previous machines. This has been achieved by improved machine movements and by the more extensive use of self-lubricating bearings and felt washers.

Attention has been given to making the cover of the machine both functionally efficient and pleasing in appearance. It is aluminium with smoothly styled lines and a distinctive two-tone grey-silver hammer finish. As part of the sound-reducing arrangements, the cover is made to enclose the machine completely and is lined with an acoustic material, easy access to the paper and ink ribbon being provided by hinged rear and window sections.

2. *Design Principles*

The teleprinter receiver is fitted with a single driving shaft, powered by a lightweight 4200-revolution-per-minute, fractional-horsepower motor. This shaft drives the selector and translator cam shafts through friction clutches. The use of a single driving shaft has the advantage of keeping distortion due to gears to a minimum, whilst the use of friction clutches ensures a constant pick-up time, which contributes to the receiver margin.

The selector cam shaft controls the selector mechanism, which converts the signal combination registered by the receiving electromagnet into a code setting on one or other of two sets of 5 pins on the translator unit. The method of selection employed secures an operating margin of ± 40 per cent at a telegraph speed of 50 bauds and ± 35 per cent at 75 bauds. An orientation adjustment is provided by a simple device that enables the selecting actions to be made earlier or later with respect to the start pulses of incoming signals by any percentage of the length of a signal element up to ± 70 per cent. This provides a simple means of checking the margin of the receiver and enables the selector mechanism to be set in the centre of the tolerance range.

The translator cam shaft controls the translator mechanism. Its main function is to convert the selecting-pin code setting into the appropriate machine action, that is into printing a character or performing a function. It has been mentioned that there are actually two sets of 5 pins. This duplication enables a combination to be set up by the selector mechanism on one set of pins at the same time that the previous combination, which has already been set up on the other set of pins, is being read by the translator. The box holding the two sets of pins is made to move forwards and backwards by the translator unit so that each set of pins engages alternately with the selector and translator mechanism. This mode of operation permits the teleprinter to print each combination immediately after selection without storing it until the receipt of the following combination.

The translator unit effects the printing of the selected characters by moving a typewheel through a simple link-type aggregate-motion mechanism (Figure 2). The typewheel is much smaller and lighter than previous Creed typewheels (see Figure 3 for a comparison between the new typewheel and that on the model-*54* teleprinter). It is mounted vertically and has 4 layers of type on it, one under the other, each layer containing 16 types. It has four degrees of movement:—

A. *Lateral movement* by a rack-and-pawl mechanism actuated by a cam in the translator unit. The typewheel traverses the length of the platen, a character at a time, except when a functional combination is received. If this combination is the carriage-return signal, the typewheel returns sharply to the beginning of the line, where it is

smoothly brought to rest by a small piston-type dashpot under the main base.

B. *Vertical movement* under dual control of the letters–figures-shift mechanism and element *2* of the received code combination. In the rest position, to which the typewheel returns after each character has been printed, the top of the typewheel is below the level of the printing point (see Figure 4), thereby ensuring immediate and complete visibility. When the translator camshaft rotates, the typewheel is raised by an amount depending on whether the selected character is in the letters or figures case and whether element *2* is a mark or space. One of 4 layers of types is thus raised level with the printing point. (The mechanical details are illustrated in Figure 2.)

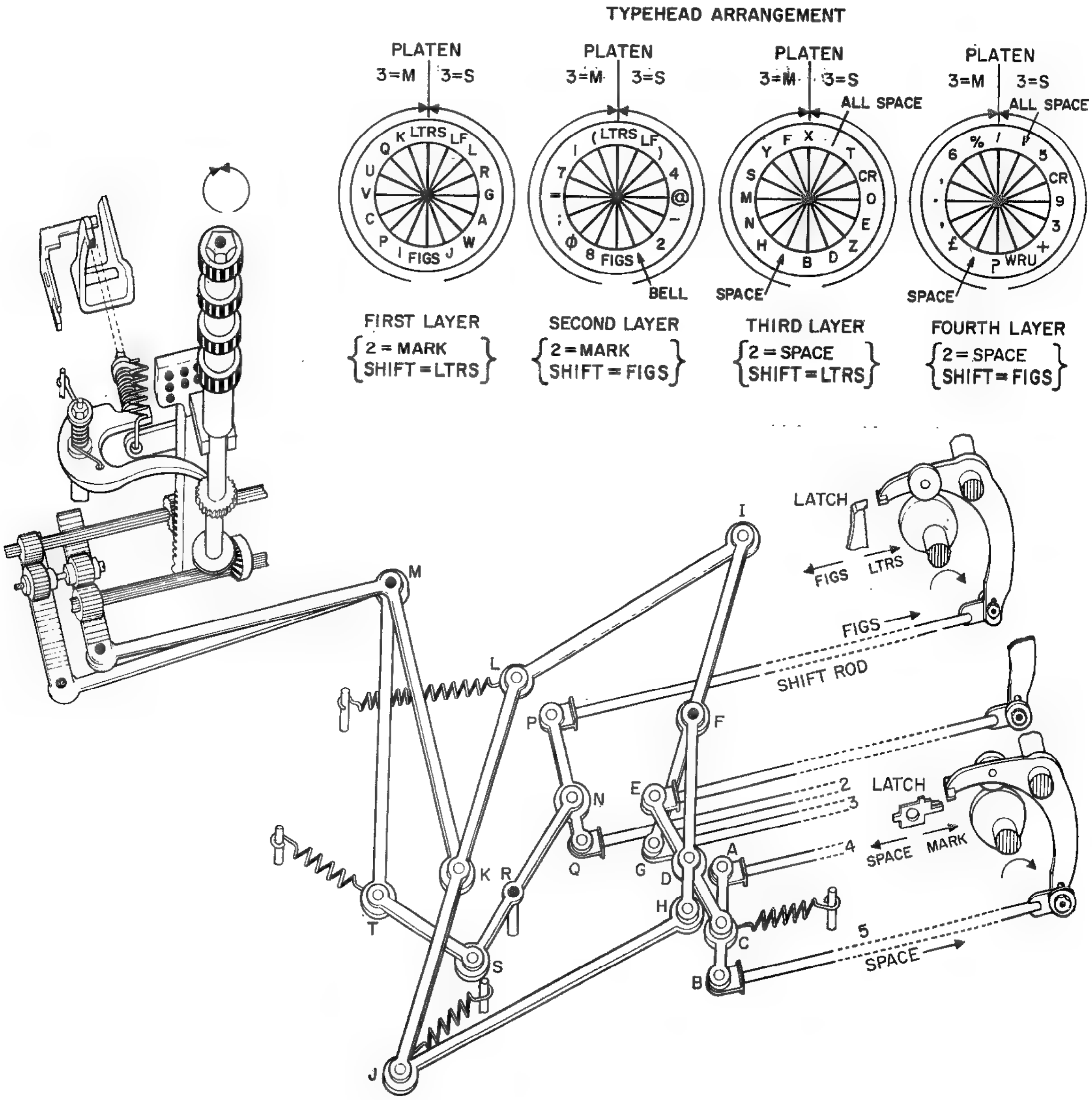


Figure 2—Aggregate-motion-mechanism diagram. Pivots *M*, *R*, and *F* are fixed. Rod movements for shift and elements *1*, *2*, *3*, and *4* are 0·5 inch (127 millimetres) and for element *5* is 0·25 inch (64 millimetres). Proportions of levers are: $AC/CB = 1/1$, $ED/DC = 1/1$, $FD/DH = 3/1$, $IF/FG = 4/3$, $LK/KJ = 1/1$, and $PN/NQ = 2/1$.

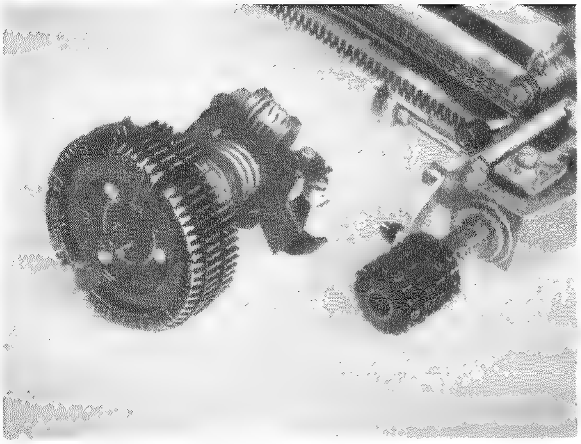


Figure 3A—Comparison of typehead units on the model Seventy-five (right) and model-*54* (left) teleprinters.

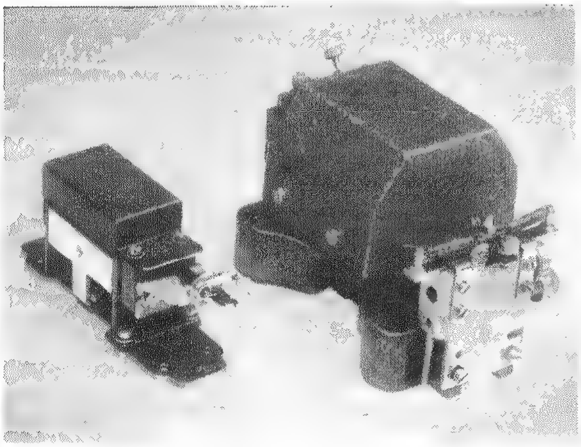


Figure 3C—Comparison of electromagnet units for the model Seventy-five (right) and model *54* (left).

C. *Rotary movement* under control of elements *1*, *3*, *4*, and *5*, of the received code combination. The positions of the selecting pins for these elements determine which of the corresponding control rods on the aggregate-motion mechanism should move to the right (see Figure 2). This determines both the direction and degree of rotation of the typewheel, the direction of rotation being controlled by element *3* and the angle of rotation, which never exceeds half a revolution, being controlled by the combined action of elements *1*, *4*, and *5*. The proportions of the levers in the aggregate-motion mechanism are so arranged that for each combination of marks and spaces for the elements *1*, *3*, *4*, and *5*, one of the

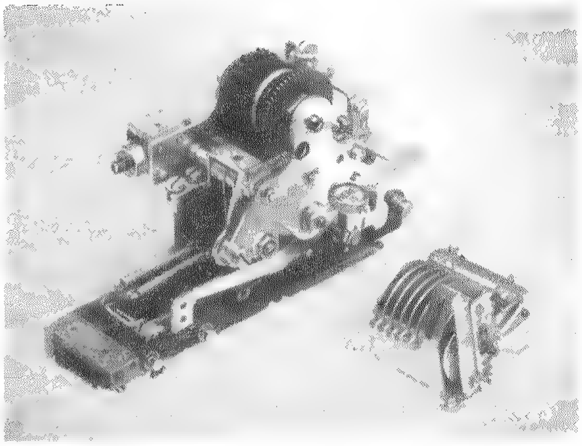
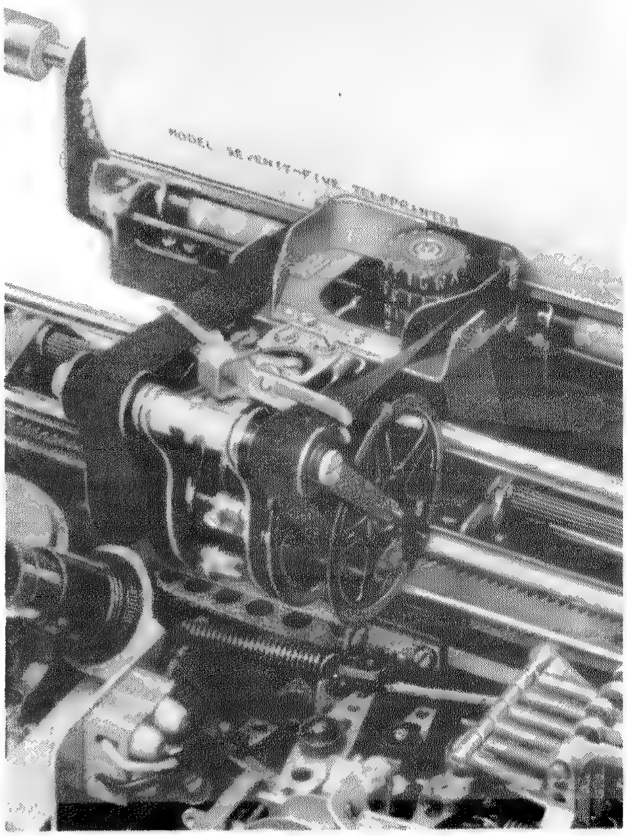


Figure 3B—Comparison of answer-back units on the model Seventy-five (right) and model *54* (left).

16 types on the selected layer is brought opposite the printing point.

D. *Forward movement* by the printing mechanism that is also energised by a cam on the translator unit. After the selected character is positioned

Figure 4—Close-up of printing point showing stationary platen and moving typehead. The typehead returns instantly at all speeds, operation being unaffected by shock, tilting, or vibration. Every character is printed immediately and is visible as soon as printed.



opposite the printing point by the aggregate-motion mechanism, the frame supporting the typewheel is swung forward. This raises the ribbon into line with the selected type and causes the typewheel to strike forward at the platen. At the end of the printing operation, the typewheel returns to its rest position and the ribbon is lowered.

The basic design feature of the translating and printing mechanisms is the use of a stationary carriage and a moving typewheel (see Figure 4). This arrangement is superior to previous Creed teleprinters (moving carriage and fixed typewheel) as it enables the width of the machine, and hence also its weight, to be substantially reduced. It also eliminates the problem of moving and smoothly arresting the mass of the paper carriage, makes the printed copy easy to read, and allows for varying the paper storing and feeding arrangements, particularly where external pre-printed sprocket-feed business stationery manifolds are employed.

The aggregate-motion mechanism and the typewheel embody a number of important design features that combine to reduce to a minimum the energy that has to be dissipated in bringing the typewheel to rest. Consequently, it has been possible to make the components associated with the typewheel smaller and to increase their life considerably. The direct method of printing that has been employed, for example, not only dispenses with the need for a typehammer and sliding types, but permits a much-smaller typewheel. Again, the aggregate-motion mechanism moves the typewheel smoothly to the printing position by the shortest route and more slowly than on previous machines.

The keyboard is of the motorized kind; that is, the power to move the combination bars is provided by the machine itself and not by the key-depressions. In this respect, the keyboard is similar to an existing Creed model used on telex teleprinters, but the touch has been improved by making it possible to depress any key well before the end of the cycle of operations initiated by depression of the preceding key. This allows the operator to type at an irregular speed about the cadence speed of the machine.

The most-novel feature of the keyboard, which makes it much simpler than other teleprinter keyboards, is the absence of a separate keyboard transmitter. When a key is depressed, the code for the key is set up on 5 combination bars. These mechanically select the code on the pins in the translator unit. The transmitter contacts, which are also mounted on the translator unit,

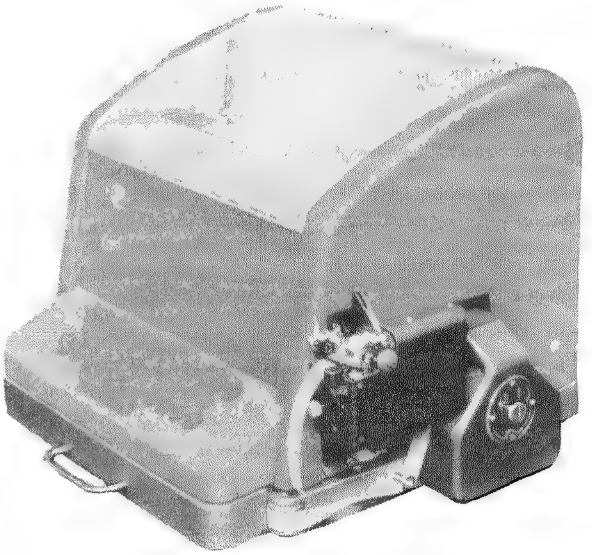


Figure 5—Close-up of reperforating attachment that fits on the right-hand side of the machine and records incoming messages in standard five-channel tape simultaneously with normal page printing. It also enables the machine to be used as a normal keyboard perforator for the origination of punched tape with coincident printed-page copy.

read off the combination from the selected pins and transmit it to line. This mode of operation avoids the need for a separate keyboard camshaft, clutch, and gears, thereby achieving simplification in design and reduction in number of parts. Also, since the combination is read off by the aggregate-motion mechanism and printed at the same time as it is transmitted to line and not, as on previous machines, after a time delay, the response of the machine to the operator's touch approximates that of a typewriter.

A consequence of using the translator camshaft for controlling keyboard transmission is that the transmission is basically $6\frac{1}{2}$ units in length; 130 milliseconds at 50 bauds. It is, therefore, necessary to insert the missing unit, and this is done

by employing a spring-controlled time-delay mechanism that extends the stop signal from $\frac{1}{2}$ unit to $1\frac{1}{2}$ units in length.

3. Further Developments

As previously mentioned, the model-Seventy-five teleprinter, while being a complete and self-contained machine, is also the basic unit of an integrated line of equipment that is being developed by the addition of special attachments or by modifying the teleprinter in various ways.

In designing the teleprinter, a great deal of attention was given to ensuring that this projected range of auxiliary machines could be developed from the parent machine with a minimum number of modifications. Such integrated design methods result in considerable advantages to customer and manufacturer alike. Smaller stocks of spare parts can be held by customers having more than one of the related types of machine, while overall development time is reduced and manufacturing economies are made possible.

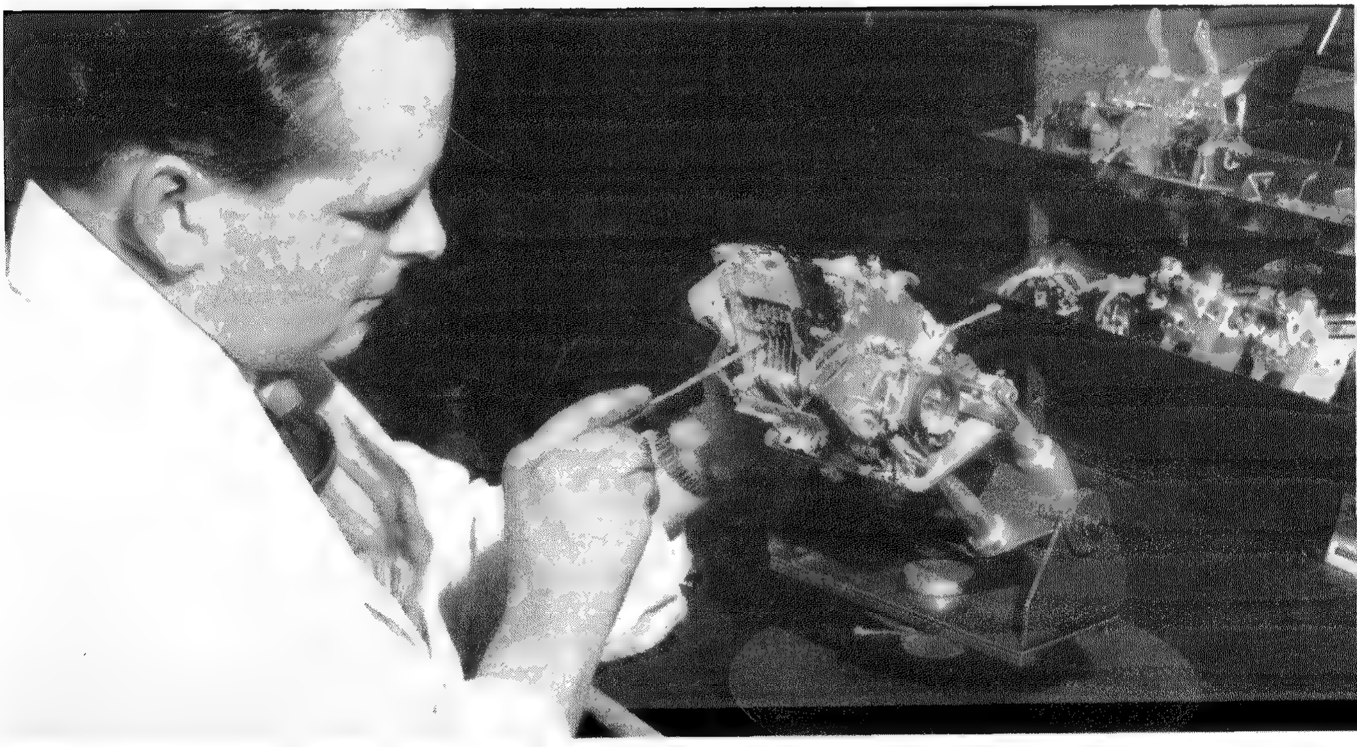
The first of the special attachments—a reperforating unit—which has recently gone into production, is illustrated in Figure 5. This fits on the right-hand side of the teleprinter, its code bars being set by the same operating levers that control the aggregate-motion-mechanism control rods. The attachment provides a perforated record of both transmitted and received signals. The punched tape, which is supplied from a drawer under the main base, issues from the attachment towards the front of the machine. Its compactness may be judged from the fact that it adds less than 3 inches (7·6 centimetres) to the width of the machine and about 1 inch (2·5 centimetres) to its height.

The other special attachments and derivative machines will include an automatic tape transmitter, a printing reperforator, and a tape teleprinter. These are, at present, being actively developed.

The mechanical local-record feature of the teleprinter enables it to be adapted easily to a variety of applications in the field of integrated data processing. One such application, which is now under development, is a teleprinter operating from 5-wire parallel input. This has been derived from the standard machine by replacing the keyboard with five electromagnets, which mechanically set the selecting pins through existing mechanisms in response to the parallel input signals.

The model Seventy-five teleprinter itself is now in full-scale production (Figure 6). Prior to this, preproduction models of the machine were given extensive trials by a number of telegraph administrations and other large organisations. These trials were most successful—how successful can be judged from the fact that orders for several thousand machines have already been received.

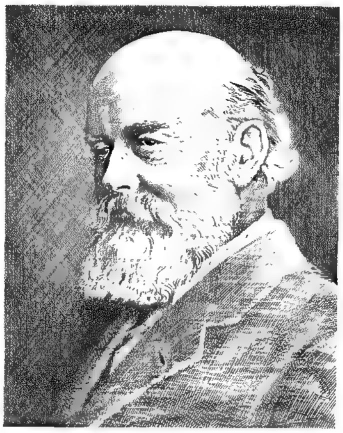
Figure 6—Assembling translator units for the model Seventy-five teleprinter.



Etching of Oliver Joseph Lodge

Oliver Joseph Lodge (1851–1940), British scientist, is depicted in the latest of the series of etchings published by the International Telecommunications Union.

Although originally planning on a business career, Lodge's interest in science led him to enter University College, London, in 1872. Starting his work as a teacher three years later, he succeeded in 1881 to the chair of physics at University College, Liverpool. In 1900, he was appointed the first principal of Birmingham University, where he continued until his retirement from academic life in 1919.



He did original work on lightning, the source of electromotive force in voltaic cells, electrolysis, ionic velocity, electromagnetic waves and wireless telegraphy, motion of the ether, and the use of electricity to disperse smoke and fog. His experiments at Liverpool with the coherer and the phenomenon of tuning, which he fully described in a patent of 1897, were important advances in radio communication. By his writings, Lodge did much to familiarize the lay public with the scientific views of his time. He also published extensively on psychical research.

Lodge was elected a Fellow of the Royal Society in 1887 and received knighthood in 1902. He was a president of the Physical Society, the British Association, the Radio Society, and the Röntgen Society The Institution of Electrical Engineers conferred on him successively its vice-presidency, honorary membership, and Faraday Medal.

The etching of Lodge is the 24th in the series that was started in 1935. On a good grade of paper measuring 9 by $6\frac{5}{8}$ inches (23 by 17 centimeters) including margins, these etchings are available at 3 Swiss francs each from Secrétariat général de l'Union internationale des Télécommunications, Palais Wilson, 52, rue des Pâquis, Genève, Suisse. The entire series is comprised of etchings of Ampère, Armstrong, Baudot, Bell, Erlang, Faraday, Ferrié, Fresnel, Gauss and Weber, Heaviside, Hertz, Hughes, Kelvin, Kirchhoff, Lodge, Lorentz, Marconi, Maxwell, Morse, Popov, Pupin, Rayleigh, Siemens, and Tesla.

High-Frequency Radio Receiver RX.5C

By L. J. HEATON-ARMSTRONG and J. D. HOLLAND

Standard Telephones and Cables Limited; London, England

DESIGNED primarily for the reception of frequency-shift telegraphy in the band from 2 to 30 megacycles per second, the *RX.5C* radio receiver incorporates a number of important features to improve operation under difficult receiving conditions.

The reception of telegraph signals in the high-frequency band is adversely affected by noise, interference, and fading, and it is usual to employ high-gain directive antennae, frequency-shift telegraphy, and dual space diversity to combat these effects.

It has been pointed out[1] that if selective fading is present, the use of the conventional limiter and discriminator for frequency-shift reception does not utilize all of the available information in the signal; some of this is destroyed in the limiter. If, for instance, selective fading is causing strong mark and very-weak space signals to be produced at the receiver input, the signal from the discriminator will consist of good mark signals followed by bursts of noise or the absence of signals, depending on the receiver gain adjustment. If, however, the signals could be observed on an oscillograph it would be immediately apparent that the noisy or missing elements should be space signals.

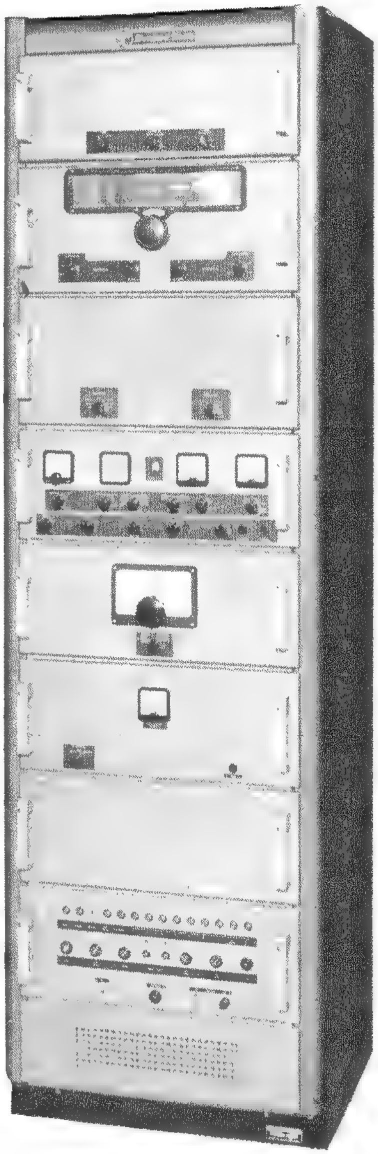


Figure 1—*RX.5C* equipment for frequency-shift telegraphy reception on the high-frequency radio band.

The receiver described here uses a technique that gives satisfactory reception provided a good mark or a good space signal alone is being received. This is equivalent to adding dual frequency diversity, and gives a decisive improvement when selective fading is present.

Another feature is the use of ratio squaring[2] for combining the dual space-diversity signals. Compared to the more-usual method of switching to the strongest signal, ratio squaring gives a worthwhile improvement in signal-to-noise ratio and eliminates trouble due to switching transients.

Another useful feature is the derivation of the voltage for automatic frequency control from both mark and space signals, so that good automatic frequency control is obtained as long as either signal is present. This mode of operation gives less residual mistuning than the use of one signal alone.

1. Equipment

Figure 1 is a photograph of the *RX.5C* receiver, which is housed in a cabinet 78 inches (198 centimetres) high, 23 inches (58 centimetres) wide, and 21 inches (53 centimetres) deep. The equipment is mounted on eight withdrawable trays and is split into small chassis units for ease of servicing and maintenance.

[1] J. W. Allnatt, E. D. J. Jones, and H. B. Law, "Frequency Diversity in the Reception of Selectively Fading Binary Frequency-Modulated Signals," *Journal of the Institution of Electrical Engineers*, Part B, volume 104, pages 98–110; March, 1957.

[2] L. R. Kahn, "Ratio Squarer," *Proceedings of the IRE*, volume 42, page 1704 (Correspondence); November, 1954.

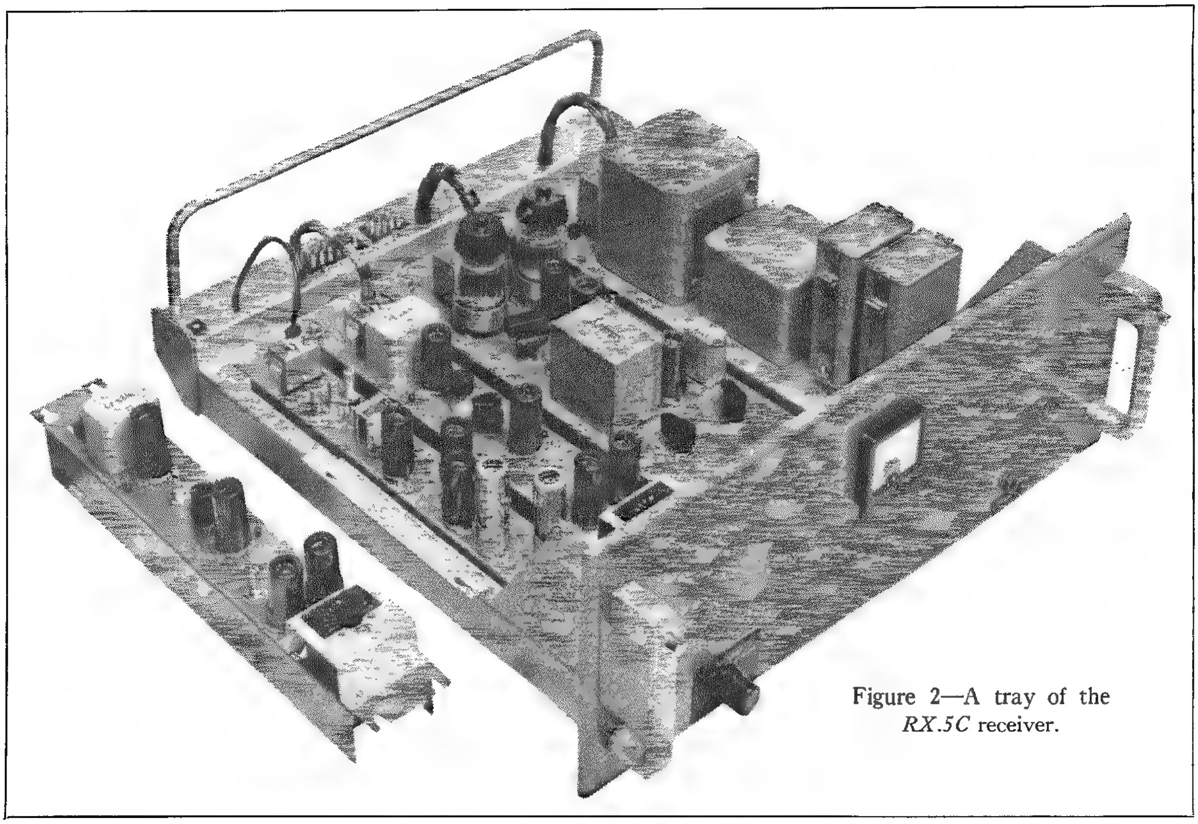
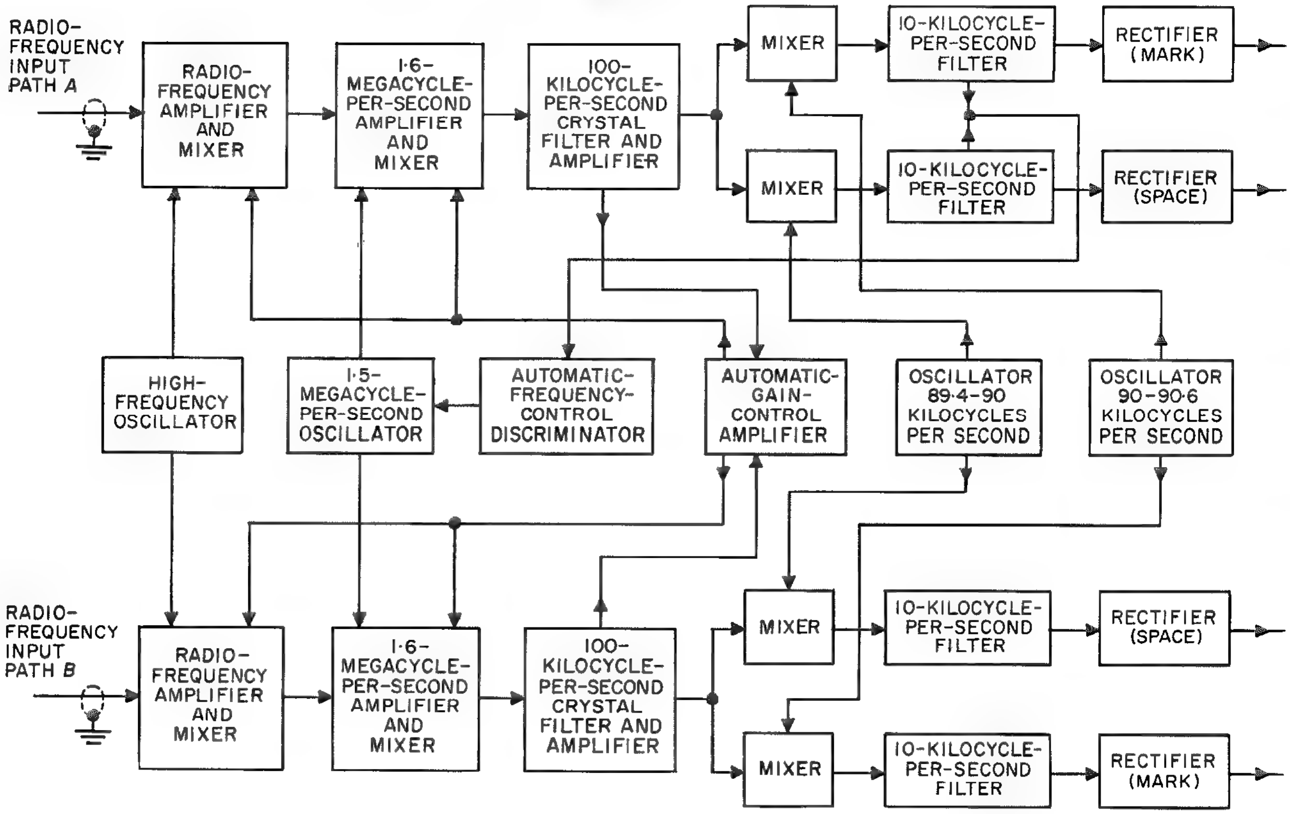


Figure 2—A tray of the *RX.5C* receiver.



A typical tray is shown in Figure 2. Front access only is required.

The equipment is suitable for operation in ambient temperatures from −20 to +45 degrees centigrade, a small blower and air filter being provided for cooling.

Silicon rectifiers and encapsulated transformers are used in the power units.

The equipment is suitable for reception of *A1*, *A2*, *A3*, *F1*, and *F6* (4-frequency diplex) signals. If all facilities are not required initially, the appropriate units can be omitted in the first instance and added later, as the cabinet is wired for them.

Only international type valves are used.

2. *Principles of Operation*

2.1 Reception of Frequency-Shift Telegraphy

Figure 3 shows a block schematic of the units used for frequency-shift telegraphy.

The receiver is of the dual space-diversity type employing triple detection. A two-stage radio-frequency amplifier covers the band from 2 to 30

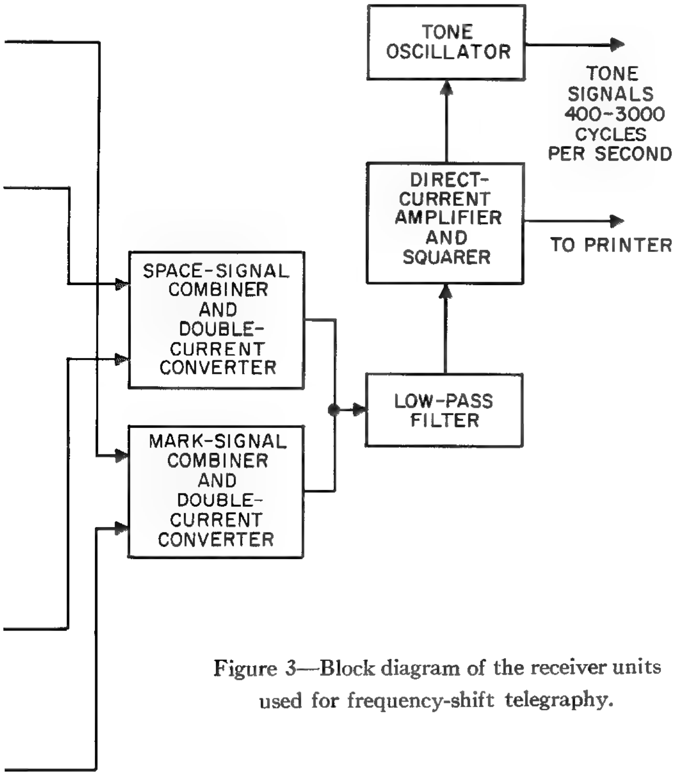


Figure 3—Block diagram of the receiver units used for frequency-shift telegraphy.

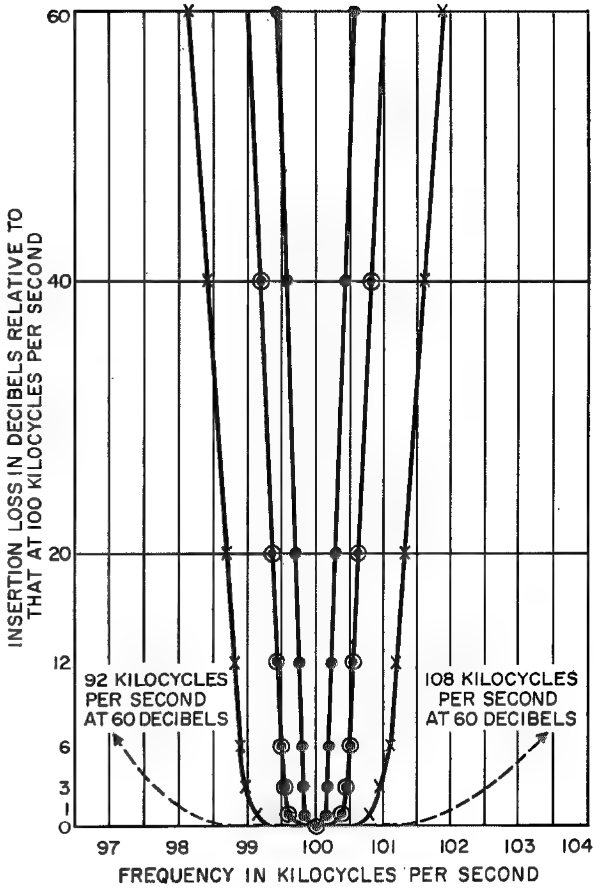


Figure 4—Response–frequency curves for crystal filters used in the 100-kilocycle-per-second section of the receiver. The nominal bandwidths are 500, 1000, 2000, and 6000 cycles per second, the latter being used for telephony.

megacycles per second. The amplifiers for each diversity path are ganged to a single control for ease of tuning. A high-frequency beating oscillator, the frequency of which can be either crystal controlled or adjustable, is used to convert the high-frequency signals to an intermediate frequency of 1·6 megacycles per second. After passing through a two-stage amplifier, the signals are then converted to a second intermediate frequency of 100 kilocycles per second and go through a crystal filter that provides the major part of the selectivity.

Filters are available with bandwidths of 500, 1000, and 2000 cycles per second, for the various signalling speeds and services.

Figure 4 shows the filter response–frequency curves. Care has been taken to obtain a good

transient response to avoid ringing, which can cause faulty operation under certain conditions of propagation.

The 100-kilocycle-per-second signals contain both mark and space frequencies, which are now separated into individual channels by applying the signals to two mixers. In one mixer, the signal is combined with a 90-kilocycle-per-second wave to convert the mark frequency to 10 kilocycles per second for selection by a filter operating at that frequency.

Similarly, an oscillation of appropriate frequency is applied to the other mixer to convert the space frequency to 10 kilocycles per second for subsequent selection by a filter. There are two 10-kilocycle-per-second outputs for the mark frequency (one from each diversity path), and these are now combined in a circuit that adds them in proportion to the square of the signal-to-noise ratios. The mark signals are then converted to double-current signals and applied to a low-pass filter. The space signals are similarly combined and converted to double-current form and applied to the low-pass filter. The signals from the low-pass filter are then applied to shaping circuits to square them before passing to the printer.

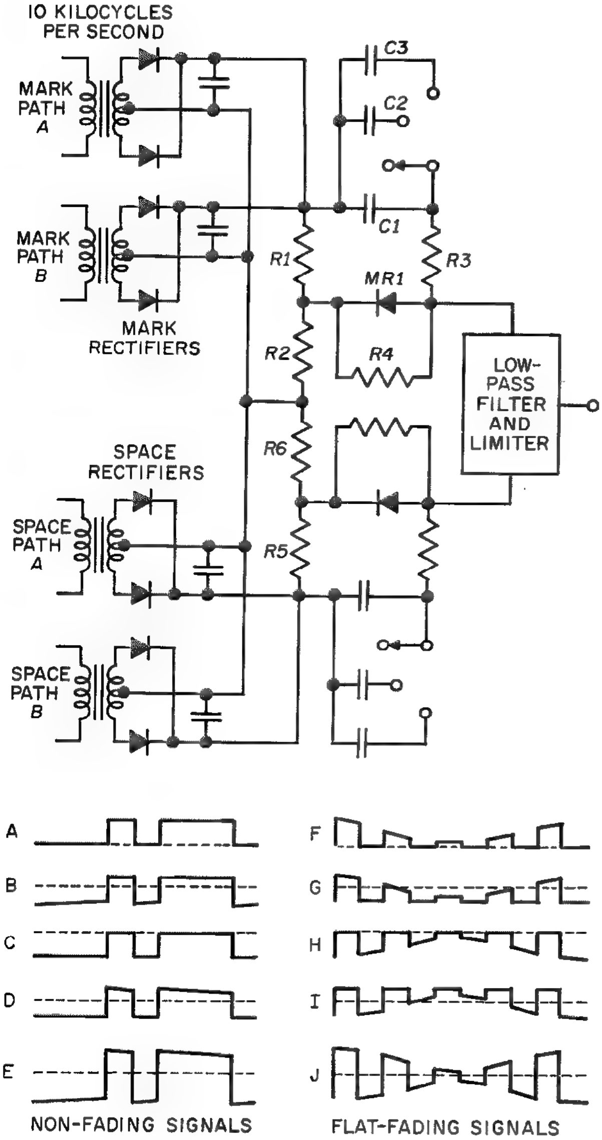


Figure 5—Details of combiner and double-current converter.

A and *F* = mark rectifier output across *R1* and *R2*.
B and *G* = mark signal at input of low-pass filter.
C and *H* = space rectifier output across *R5* and *R6*.
D and *I* = space signal at input to low-pass filter.
E and *J* = combined mark and space signals at input to low-pass filter.

2.2. Combiner and Double-Current Converter

It has been shown by Kahn[2] that for optimum results the signals from the two diversity paths should be combined in proportion to the square of their signal-to-noise ratios. A close approximation to ratio squaring can be obtained very simply[3] by connecting the two signals to a common resistance and choosing the source impedance equal to $R(2^{1/2}\text{-}1)$, where R is the common resistance.

Figure 5 shows the circuit for the combiner and double-current converter. The mark signals from the dual paths are rectified and combined in a common load, the value of which is mainly determined by the two equal resistances *R1* and *R2*, which are smaller than *R3* and *R4*. Half the voltage produced by the mark signal appears across *R2* and is applied to the filter input. The other half appears across *R1* and charges *C1*. When the mark signal ceases, *C1* discharges through *R1*, *R3*, and *R4*, but *R4* is large compared to *R1* and *R3*, so that nearly all the voltage from *C1* appears across *R4* and is applied to the filter as a negative voltage. The mark signal has thus been converted to a double-current signal shown at 5B. The space signals are similarly

[3] R. T. Adams, British Patent 801 165.

treated (5C and 5D) and added in series with the mark signals to give a combined signal output shown in 5E.

The time constants for charge and discharge of *C1* are chosen so that *C1* charges in the time of the shortest signal element and discharges in a somewhat longer time than that of the longest signal element.

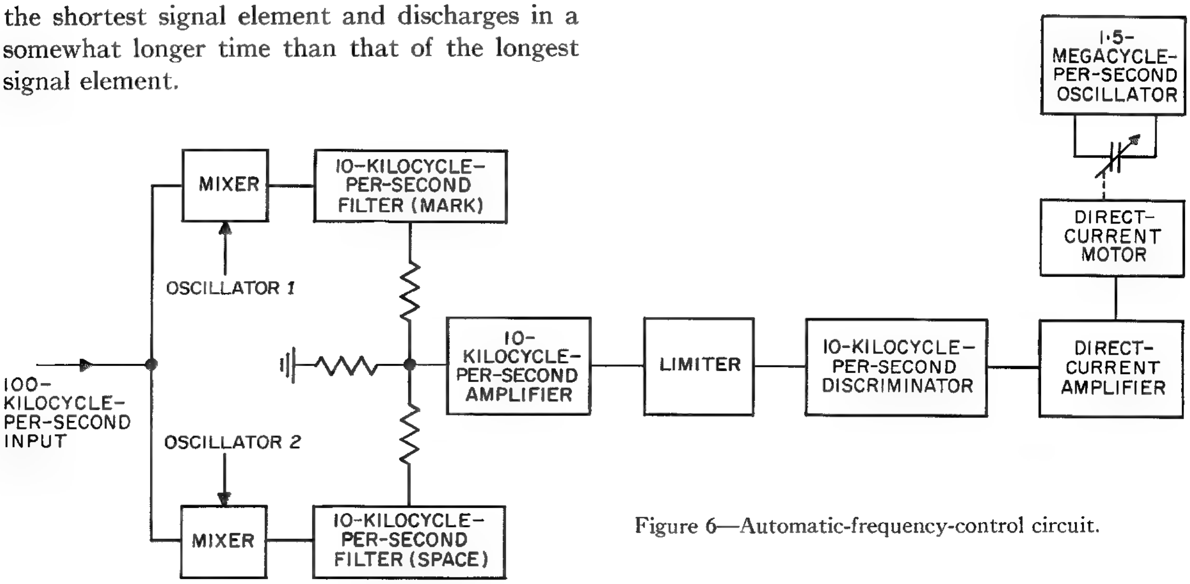


Figure 6—Automatic-frequency-control circuit.

If either signal is present the printer will operate; there is thus a great advantage over the conventional limiter–discriminator, which is almost certain to give errors if the mark or space signal only is present.

It should also be noted that a continuous mark or space signal will be transmitted through the combiner. This is important for teleprinter operation.

The effect of fading will now be considered. If fading is slower than the time constant of $C1 \times R4$, the system will operate as described above. If the fading period is less than the time constant of $C1 \times R4$ but is of the nonselective type, that is, both mark and space frequencies are present, the waveform will remain symmetrical as shown in Figure 5E, and will merely vary in amplitude as shown in Figure 5J. No errors will be produced provided the signal-to-noise ratio is high enough.

Under conditions of fast selective fading that has a period appreciably less than the time constant $C1 \times R4$, the waveform will become unsymmetrical and errors will be produced. This condition is shown in Figure 5G, which shows mark signals only, the space signals being presumed absent due to selective fading. This condition will very rarely occur in practice and, if present, can be avoided by decreasing the time constant. The circuit will then behave like a normal discriminator and will operate satisfactorily provided there is a sufficiently good signal-to-noise ratio on both mark and space signals.

2.3. Automatic Frequency Control

Figure 6 shows the automatic-frequency-control circuit. The 10-kilocycle-per-second mark and space signals from one path are combined, amplified, and limited, and then applied to a discriminator. The direct-current output of the discriminator is amplified and used to drive a direct-current motor that is coupled to a small trimmer capacitor connected across the 1·5-megacycle-per-second oscillator. When used on twinplex the automatic-frequency-control system operates from the four signalling frequencies, thus giving a control voltage at all times.

2.4 Twinplex, Four–Frequency Diplex

Twinplex[4] is a system by which two communication channels can be maintained simultaneously using a transmitter that need not be of the

[4] C. Buff, "Twinplex and Twinmode Radiotelegraph Systems," *Electrical Communication*, volume 29, pages 20–33; March, 1952.

linear amplifier type. Frequency-shift keying is employed using four frequencies, only one of which is present at any instant. A typical code combination is shown in Table 1. The frequency shift is normally 600 or 1200 cycles per second. *F1*, *F2*, *F3*, and *F4* are the four signalling frequencies.

Figure 7 shows the receiver circuits for one diversity path. The signals at 100 kilocycles per second go to four mixers that are also supplied by oscillators *1* through *4* having frequencies that will convert each of the four signalling frequencies to 10 kilocycles per second. The four frequencies are thus separated and then rectified and applied to the load to produce a positive or negative voltage as required. Several code combinations are in use; a switch is therefore provided in the receiver to connect the rectifier outputs to either side of the load, and thus make up any combination.

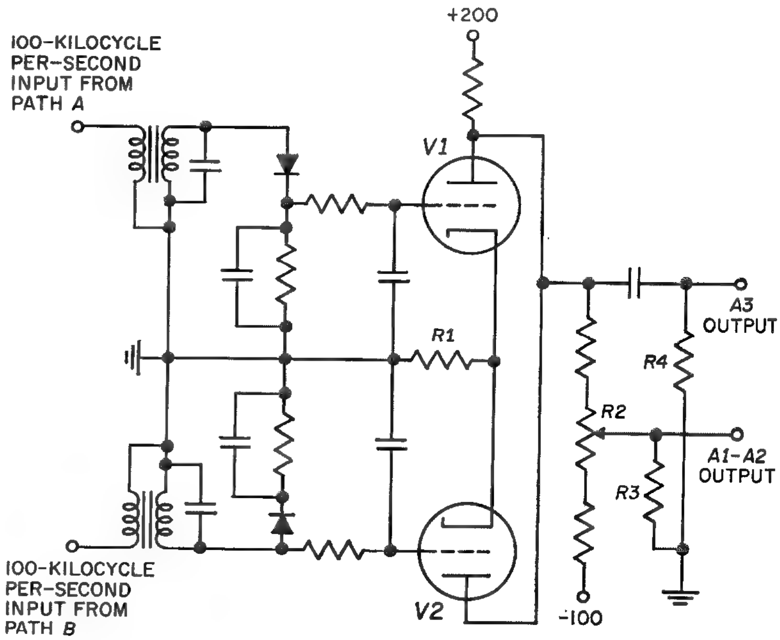


Figure 7—Combiner for four-frequency diplex.

The second diversity path is identical with that just described. Combining is done by paralleling the rectifiers. By a correct choice of the value of load resistance and rectifier source resistance, a ratio-squaring characteristic is obtained.

2.5 *A1*, *A2*, and *A3* Signals

Outputs taken from the crystal filters at 100 kilocycles per second are impressed on the combining circuit shown in Figure 8. For all modes of working the signal envelopes go to the combiner valves *V1* and *V2*. On telephony the combined audio-frequency components are developed across *R4*, and on telegraphy the keyed components appear across *R3*.

TABLE 1

Twinplex Signalling Frequencies

Condition of: Channel *1*	Condition of: Channel *2*	Designated Frequency	Deviation from Carrier Frequency in Cycles per Second
Mark	Mark	*F4*	+600
Mark	Space	*F3*	+200
Space	Mark	*F2*	−200
Space	Space	*F1*	−600

The method of combining provides for attenuation of the weaker signal due to the common cathode resistance *R1* and can be described as a linear attenuating and adding system; it is approximately 0·5-decibel inferior to the ratio-squaring system.[2]

On telegraphy, the combiner provides a double-current output from single-current inputs, and this allows the direct-current amplifier and squaring circuits shown in Figure 3 to be keyed effectively as on *F1* working.

This conversion process is obtained by the setting of the threshold bias control *R2*, which is adjusted to give a positive-going output across *R3* on space (no-signal condition) and a negative-going output on mark. The setting of *R2* provides a means for controlling systematic distortion.

3. Monitoring Circuits

Facilities are provided for monitoring the following:

A. Direct- and alternating-current levels.

B. Signal and output levels.

C. Frequency.

The first item is covered by monitoring the cathode currents of all valves in conjunction with a wander lead and built-in monitor meter. Each tray carries a socket that is wired to a tray tap switch and meter, and insertion of the wander plug into lead-through pins situated at each valve provides a measurement of valve performance. This method is relatively inexpensive and less complex than one in which each valve is wired to a tap switch. The meter is also, in conjunction with rectifiers, calibrated to indicate the tone output to line.

Item *B* is covered by a monitor meter that indicates the field strengths of paths *A* and *B* in terms of level at the receiver input. The instantaneous signal levels at the 10-kilocycle-per-second branched outputs can be observed on electronic tuning indicators. Separate centre-zero meters are included in the channel-*1* and -*2* telegraph outputs. These meters can be switched to indicate the currents through locally connected teleprinters or a remote load.

The output telegraph level must be free from systematic distortion, and this is obtained by injecting a level at 50 cycles per second into the input terminals of the bistable circuits used on *A1* and *F1* working and adjusting the line output currents for zero indication in centre-zero meters. This procedure checks all the circuits after the demodulating process.

Circuits between the antenna and the various demodulators can be checked, with no input signal, by measurement of input thermal noise in the field-strength meters.

The last item, frequency determination, is accomplished by the use of three crystal-controlled oscillators at 250, 100, and 90 kilocycles per second arranged in various circuit configurations for monitoring purposes as follows:—

A. The 250- and 100-kilocycle-per-second oscillators serve as reference check points for the adjustable-frequency first beating oscillator.

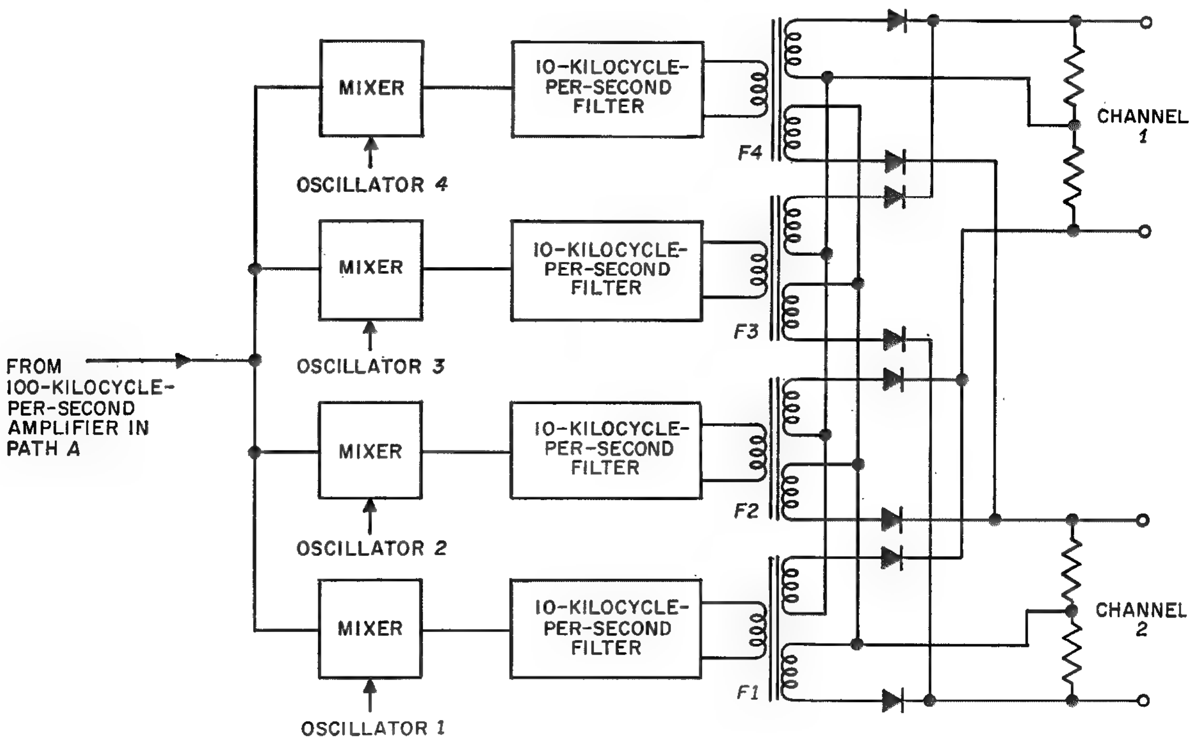


Figure 8—Combiners for telegraphy (*A1* and *A2*) and telephony (*A3*).

B. The 100-kilocycle-per-second oscillator is used to check the 100-kilocycle-per-second discriminator and, by beating with the 90-kilocycle-per-second crystal, yields a 10-kilocycle-per-second output for checking the automatic-frequency-control discriminator and amplifier circuits that operate at 10 kilocycles per second.

C. The 90-kilocycle-per-second crystal oscillator, by substitution, checks the high-stability frequency-shift oscillators shown in Figure 3, which operate at 89·4 to 90 and at 90 to 90·6 kilocycles per second. The shift oscillators are ganged together and are calibrated in shift up to 1200 kilocycles per second. These oscillators, after checking, can be switched into the circuit in turn for a functional check of the motor circuits of the automatic frequency control.

All the main supply circuits are protected by miniature circuit breakers instead of fuses. These are mounted on the front panel of the power unit and provide rapid visual monitoring of breakdown due to overload. These breakers are designed in such a manner that the circuits cannot be re-established until the fault is cleared.

4. *Typical Performance*

Frequency Range	2 to 30 megacycles per second in 7 bands
Input Impedance	75 ohms, coaxial transmission line
Noise Factor	6 decibels at 28 megacycles per second 4 decibels at 4 megacycles per second
Sensitivity	0·15 microvolt in series with 75 ohms required at 28 megacycles per second on frequency-shift operation using 1000-cycle-per-second pass-band and 400-cycle-per-second shift for 1 error per 1000 characters.
Image Suppression	Greater than 80 decibels
Intermediate-Frequency Breakthrough at 2 Megacycles per Second	Greater than 70 decibels
Selectivity at First Intermediate Frequency	12 kilocycles per second at 3-decibel attenuation 34 kilocycles per second at 20-decibel attenuation
Selectivity at Second Intermediate Frequency	
With 0·5-Kilocycle-per-Second Filter	0·4 kilocycle per second wide at 3 decibels of attenuation 1·1 kilocycles per second wide at 60 decibels of attenuation
With 1-Kilocycle-per-Second Filter	0·9 kilocycle per second wide at 3 decibels of attenuation 2 kilocycles per second wide at 60 decibels of attenuation
With 2-Kilocycle-per-Second Filter	2 kilocycles per second wide at 3 decibels of attenuation 3·8 kilocycles per second wide at 60 decibels of attenuation
With 5-Kilocycle-per-Second Filter	5 kilocycles per second wide at 3 decibels of attenuation 16 kilocycles per second wide at 60 decibels of attenuation
Distortion	Less than 5 per cent at 50 bauds for failure of mark or space signal on frequency-shift operation
Frequency Stability	
First Beating Oscillator (Crystal Controlled)	$\pm$ 50 parts per million for variation of $\pm$ 20 degrees centigrade about a mean ambient of 25 degrees centigrade and $\pm$ 5-percent variation in supply voltage.
Adjustable-Frequency Oscillator	$\pm$200 parts per million for temperature and supply variations specified above
Second Beating Oscillator	$\pm$200 parts per million for temperature and supply variations specified above

Frequency-Shift Oscillator (90 Kilocycles per Second)	±40 parts per million for variation in temperature and supply voltages specified above
Blocking	The unwanted signal must be at least 45 decibels above the wanted signal if it is 5 kilocycles per second off tune
Automatic Gain Control	Less than 6 decibels rise in output level for input variation from −12 to +80 decibels relative to 1 microvolt
Automatic-Frequency-Control Residual Mistune	Will follow drifts up to ±3 kilocycles per second with residual error of less than 5 cycles per second at 50 bauds or 20 cycles per second at 200 bauds
Capture Level	Synchronism with the wanted signal is not lost if the unwanted signal is at the same frequency and the level is not greater than −6 decibels relative to the wanted signal
Keying Speeds	Up to 200 bauds
Output Levels	
Telegraphy	50–0–50 milliamperes direct current. Tone output +10 decibels relative to 1 millwatt in 600 ohms
Telephony	+10 decibels relative to 1 millwatt in 600 ohms
Power Consumption	Not exceeding 600 volt–amperes with all units operating

Recent Telecommunication Development

Electronic Avigation Engineering

INTERNATIONAL Telephone and Telegraph Corporation is the publisher of a recently released book entitled Electronic Avigation Engineering, which was written by P. C. Sandretto, vice president and technical director of ITT Laboratories.

The book is divided into 4 parts and 17 chapters. A short introduction is given for each part. The part and chapter titles are as follows.

Part A—En-Route Long-Distance Zone

Chapter 1—Airborne Direction Finders and Radiophares
Chapter 2—Four-Course Low-Frequency Radio Range and Markers
Chapter 3—Consol
Chapter 4—Some Low-Frequency Developments
Chapter 5—High-Frequency Direction Finding from Ground Stations
Chapter 6—Loran
Chapter 7—Electronic Pilotage
Chapter 8—Electronic Aids to Dead Reckoning

Part B—En-Route Short-Distance Zone

Chapter 9—Very-High-Frequency Phase-Comparison Omnidirectional Radio Range
Chapter 10—Distance-Measuring Equipment
Chapter 11—Some Avigational Aids for the Short-Distance En-Route Zone
Chapter 12—Tacan

Part C—Approach and Landing Zone

Chapter 13—Airport Surveillance Radar
Chapter 14—Fixed-Beam Low-Approach Systems
Chapter 15—Radar Low-Approach Systems
Chapter 16—Landing Altimetry

Part D—Airport Zone

Chapter 17—Airport Surface Detection Equipment.

The book is 6 inches (15 centimeters) by 9 inches (23 centimeters) and contains 775 pages of text, 16 pages of index, 527 figures, 667 equations, and 380 references in selected bibliographies. It is available postpaid at $9.50 per copy or at $7.60 per book in lots of 12 or more to a single address from International Telephone and Telegraph Corporation, Technical Publications Section, 67 Broad Street, New York 4, New York.

Direct-Printing Receiving Systems at Low Radio Frequencies

By L. J. HEATON–ARMSTRONG and J. D. HOLLAND

Standard Telephones and Cables, Limited; London, England

BEFORE 1954 little information was published on the use of direct-printing facsimile and telegraph operation with transmission in the low radio-frequency band. In that year Doutre made a preliminary study of some of the factors that affected the reliability, cost, and complexity of a ground-to-air broadcast service intended for direct printing of weather information in aircraft flying over the North Atlantic.

Doutre estimated that, on a frequency of approximately 100 kilocycles per second, 10 kilowatts of power radiated from each of two stations located in coastal areas on both sides of the Atlantic would provide adequate coverage over the entire route with a measure of overlap. His figure appears to be a realistic estimate from an analysis of the flight tests that have been made.

Increasing interest in these frequencies has been shown by other users, mainly for newscast services and for the reception of meteorological information by weather ships. This has led to the development of narrow-band frequency-shift receivers designed for unattended operation in the band from 90 to 130 kilocycles per second.

For ground-to-air service the receiver is coded *SR.24*. It uses 11 valves and has a built-in rotary transformer operating from a 24-volt direct-current supply. This model will shortly be replaced by a design employing transistors and powered from a 115-volt 400-cycle-per-second supply. The power consumption of this model is less than one-tenth that of the *SR.24*. This receiver becomes an integral part of the model Seventy-five teleprinter supplied by Creed & Company.

For use at sea or for press work a receiver coded *RV.14* has been developed. This is similar to the *SR.24* but operates from alternating-current mains. All these receivers give a choice of 4 crystal-controlled frequencies.

1. *Choice of Bandwidth and Shift*

The bandwidth occupied by the transmission should be as narrow as possible to avoid interference with other services and also to enable the receiver bandwidth to be a minimum, thus giving the best possible signal-to-noise ratio at the detector.

The modulation index is $m = 2D/B$ where $2D$ is the frequency shift in cycles per second and B is the keying speed in bauds. The relative amplitudes of the carrier and significant sidebands are shown in Table 1 for various values of m. The figures apply to square-wave keying with the higher-frequency components removed.

TABLE 1

RELATIVE AMPLITUDES OF CARRIER AND SIDEBANDS

m	Amplitude of Carrier	Amplitude of Sidebands			
		1st	2nd	3rd	4th
0·5	0·87	0·28	0·75	0·03	0·015
1·0	0·68	0·49	0·2	0	0·03
1·3	0·425	0·53	0·3	0·05	0·03
1·5	0·28	0·53	0·37	0·1	0·03
2·0	0	0·4	0·52	0·22	0·01
2·4	0·15	0·25	0·34	0·53	0·1

The receiver bandwidth must be sufficient to include all significant components, and an allowance for transmitter and receiver frequency instabilities must also be made. The transmitter instability will be approximately ±2 cycles per second. The receiver oscillator instability will be ±2 cycles per second and the intermediate-frequency-discriminator instability will be ±5 cycles per second making a total of ±7 cycles per second for the receiver or ±9 cycles per second for the system. Adding this to the bandwidth required

TABLE 2

BANDWIDTH FOR LOSS OF 1 DECIBEL AT KEYING RATE OF 45·5 BAUDS

m	Bandwith in Cycles per Second	To Include Sideband
0–0·5	63·5	1
0·5–1·5	109	2
1·5–2·4	154·5	3

to take in the significant sidebands, when $B =$ 45·5 bauds, gives the values shown in Table 2.

Values of m below 0·5 will give poor efficiency because nearly all the power will be in the carrier.

The best value of m is about 1·3 to 1·5 as the bandwidth is then the narrowest possible consistent with a small carrier power. The best value of shift is therefore 59 to 68 cycles per second for a 45·5-baud signalling speed.

2. *Antennas*

The receivers have to accommodate a wide range of open-wire capacitive antennas, and provision is also made for the use of omnidirectional loops.

The balanced double-loop type of antenna has two main advantages in aircraft; namely, reduction in drag and good anti-static properties. This applies particularly to the flush-mounting type since the surface is protected by the boundary layer on the surface of the aircraft. The electrostatic charges will be small, and can be conducted away by graphite impregnation of the loop housing.

deteriorated by a factor depending on the ratio of tuning capacitance to antenna capacitance, and only becomes equal to the loop antenna when the antenna capacitance alone forms the total tuning capacitance; assuming equal values of Q for the loop and the inductance forming part of the tuned circuit associated with the open antenna. For this reason, with open antennas, the stray capacitance must be kept as low as possible, and any impedance-matching device must be mounted as close to the antenna as possible: this condition is not easy to obtain on some installations.

An omnidirectional pattern within ±2 decibels can be obtained from two crossed loops providing the coils are coupled with a reactive mutual impedance to obtain a 90-degree phase shift between the contributions of the loops before combination. The optimum coupling impedance is that which produces critical coupling, and this can give rise to circuit complexity if the frequency band to be covered is greater than about 1·3 to 1; necessitating adjustment of the reactive elements in discrete steps.

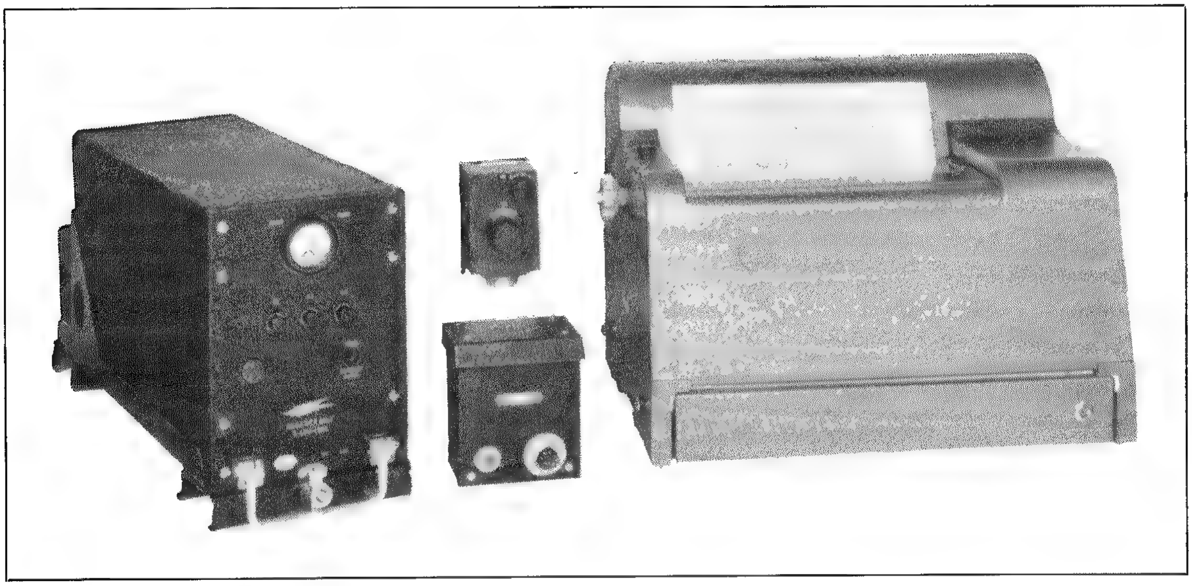


Figure 1—Equipment for a receiving installation.

Comparison of the pick-up factor between given dimensions of the open-wire antenna and the loop antenna are misleading if based on consideration of effective height alone. The signal-to-noise performance of the open antenna is

Recent promising experiments with suppressed antennas may provide an alternative solution to the use of balanced loops or open-wire antennas for aircraft use. For example, a plate having an area of 2·5 square feet (0·23 square metre)

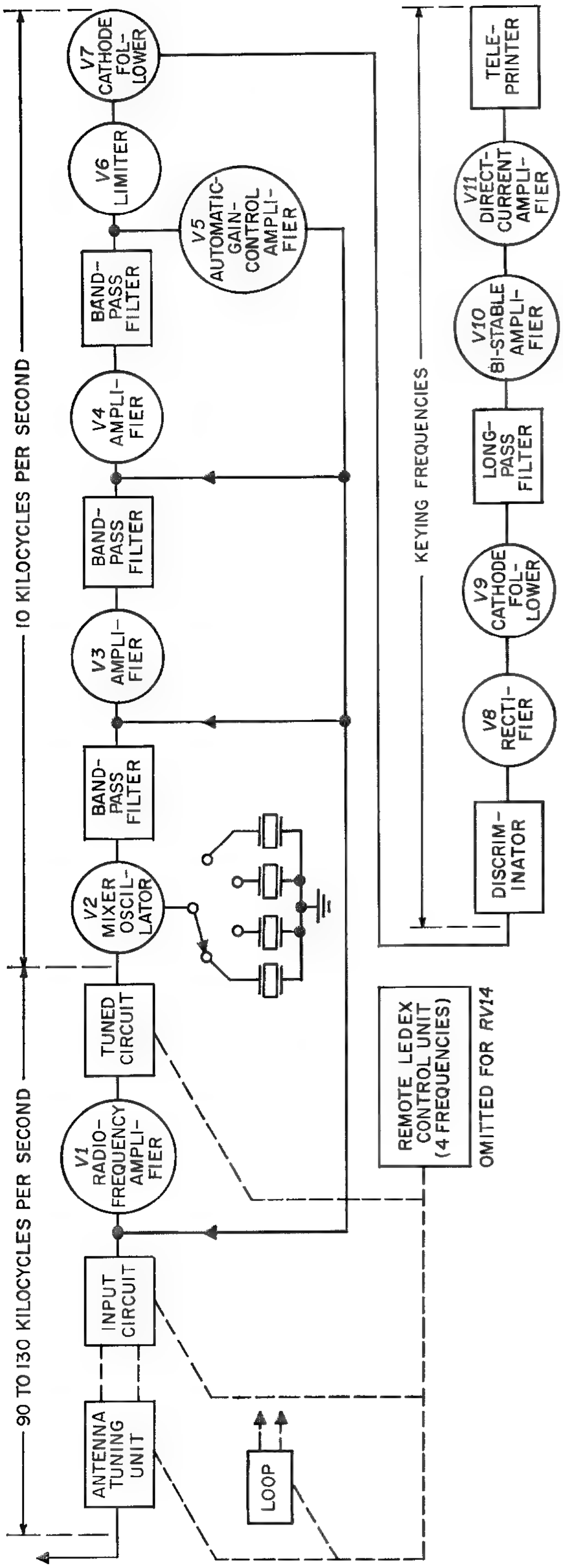


Figure 2—Block diagram of the *SR.24* and *RV.14* receivers. Power for the *SR.24* is obtained from a rotary transformer and for the *RV.14* from the alternating-current mains.

mounted 2 inches (5 centimetres) off the aircraft chassis has the same pick-up factor as an open-wire antenna some 6 feet (1·8 metres) in length and mounted 8 inches (20 centimetres) away from the fuselage. The capacitance of the open type is about 25 picofarads and that of the plate about 80 picofarads.

A plate mounted under and to the rear of the aircraft should have good immunity from rain static.

For electrically resonated antennas using ferrite rods, the best type of rod is one in which the product of effective permeability and the Q of the coil has a maximum value.

An improved pickup factor can be obtained by use of a number of rods with the windings suitably connected and with minimum magnetic coupling among them.

3. *SR.24 Receiver*

The *SR.24* receiver was designed for ground-to-air use. A photograph of the equipment for an installation is shown in Figure 1. A block diagram is given in Figure 2.

A band-pass filter is provided before the first amplifier to minimize the effects of strong unwanted signals in coastal areas.

The tuning and crystal circuits are remotely controlled by a Ledex mechanism to provide choice of four frequencies. The crystals have been specially developed for this application. Their dimensions are approximately $1\frac{1}{4}$ by $1\frac{1}{16}$ by $\frac{3}{8}$ inch (32 by 27 by 9·5 millimetres). The maximum frequency variation between -20 and $+55$ degrees centigrade does not exceed ± 42 parts per million. The crystals can be set to the nominal frequency by circuit adjustment.

A total of three critically coupled resonators are used at an intermediate frequency of 10 kilocycles per second. The circuits are temperature compensated, and the bandwidth at 3 decibels down varies between 115 and 135 cycles per second and between 415 and 450 cycles per second at 60 decibels down for a change of 55 degrees centigrade.

The circuits following the discriminator are direct-current coupled and the bi-stable amplifier is keyed at the half-amplitude of the demodulated wave. This represents a compromise between coincident noise voltages occurring at the signal transitions and noise arising during the steady-

state condition of the wave. For a shift of 40 cycles per second the maximum degree of frequency error that can be tolerated is therefore ±10 cycles per second.

The discriminator inductances are temperature compensated and encapsulated. The frequency drift of the discriminator due to the combined effects of temperature variation of 55 degrees centigrade and humidity variations up to 75 percent relative humidity does not exceed 8 cycles per second.

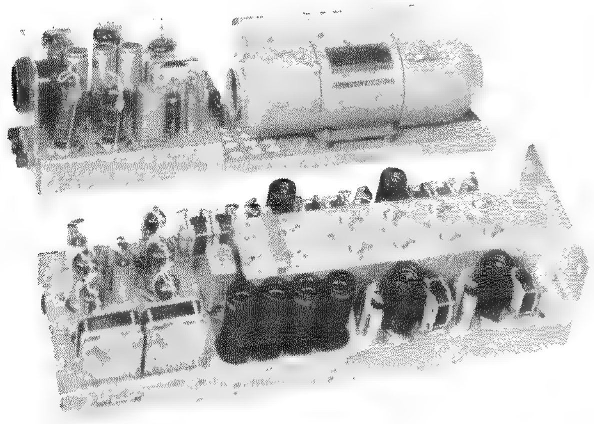


Figure 3—The *SR.24* receiver and its power supply are mounted on two chassis.

The signal-to-noise ratio is approximately 30 decibels for an input of 1 microvolt into an artificial antenna of 50 picofarads in series with 10 ohms for a bandwidth of 100 cycles per second.

Single-current open-circuit working is used for the model Seventy-five teleprinter, that is, zero magnet current on mark and about 40 milliamperes on space. The performance of the receiver is maintained for variations in the input direct-current supply of ±2 volts of a nominal supply of 28 volts.

The complete installation comprises five units, which are described below. Their dimensions, including allowances for all projections, are given in Table 3.

3.1 Receiver and Power Unit

The units that make up the receiver are shown in Figure 3. One unit contains all the circuits from the antenna to the low-pass filter, and the other unit contains a rotary transformer, stabilizing valves for the high-tension and bias supplies, and the keying circuits. Provision is made for monitoring the field strength via a jack on the front panel.

3.2 Antenna Coupling Unit

The antenna coupling unit contains one circuit coupled via a twin coaxial cable to the receiver input circuit and forms a band-pass filter covering the range from 90 to 130 kilocycles per second. The four pre-set trimmers, corresponding to a choice of four frequencies in the band, are switched by a Ledex mechanism operated from the control box.

3.3 Control Unit

A power-supply switch, four-position switch for frequency control, and an indicator lamp are included in the control unit.

3.4 Backplate Junction Box

The junction box is mounted in the tray immediately behind the receiver and power unit.

3.5 Teleprinter

The model Seventy-five teleprinter is described in full in another paper in this issue and will not be treated here.

TABLE 3

Dimensions of Receiving Units

Unit	Height in Inches (Millimetres)	Width in Inches (Millimetres)	Depth in Inches (Millimetres)	Weight in Pounds (Kilograms)
Receiver–Power	7·875 (200·0)	5·875 (149·2)	17·375 (441·3)	17 (7·71)
Antenna Coupling	5·518 (140·2)	4·986 (126·6)	2·749 (69·8)	1·56 (0·71)
Control	3·687 (93·7)	2·280 (57·9)	3·062 (77·8)	0·75 (0·34)
Junction Box	4·5 (114·3)	5·796 (147·2)	3·312 (84·1)	1·5 (0·68)
Teleprinter	14·625 (371·5)	15·875 (403·2)	14·0 (355·6)	35 (15·88)

4. *RV.14* Receiver

A photograph of the *RV.14* receiver is shown in Figure 4. This receiver uses the same circuits as the *SR.24* receiver except for the following modifications:—

A. Double-current output at 30–0–30 milliamperes.

B. Power is supplied from the 50-cycle-per-second mains at 110 or 240 volts.

C. Systematic bias distortion can be reduced to a negligible degree by a control brought out to the front panel. This control is coupled to one of the discriminator tuning capacitors and can be operated with reference to a centre-reading meter on the panel.

D. A meter is provided for monitoring field strength.

E. Provision is made to include muting circuits to prevent the printing of random symbols during conditions of high noise or interference.

The *RV.14* receiver is $12\frac{1}{8}$ inches (308 millimetres) high, $22\frac{1}{2}$ inches (572 millimetres) wide, and $14\frac{1}{4}$ inches (362 millimetres) deep.

5. *Transistor Receiver*

A receiver using transistors has been developed to replace the *SR.24*. A photograph of it is shown in Figure 5 and a block diagram in Figure 6.

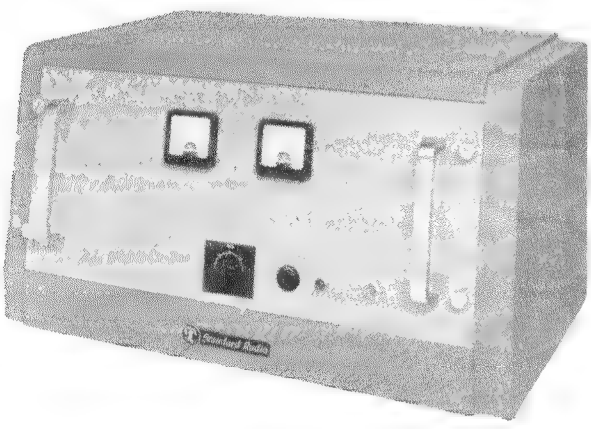


Figure 4—The *RV.14* receiver.

The receiver is mounted in a cast plinth 2 inches (51 millimetres) deep that follows the contours of the model Seventy-five machine. Normally, this space is reserved for a tape reel box for versions of the machine requiring a tape-perforating attachment. This is not required for the airborne service. Installation is completed by connection to an antenna and a 400-cycle-per-second single-phase power supply at 115 volts.

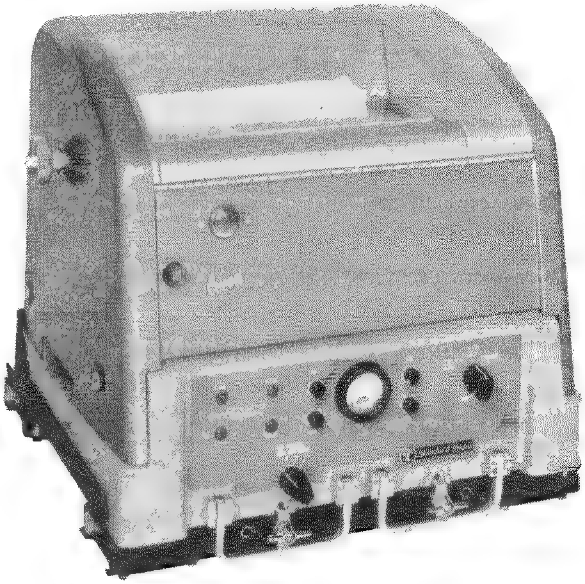


Figure 5—A receiver using transistors is mounted in the case of the model Seventy-five teleprinter. It utilizes the space in which a tape reel box would normally be accommodated.

Two critically coupled encapsulated band-pass filters are used in the radio-frequency stage, and provision is made for a loop antenna.

The inherent internal feedback of the transistor radio-frequency amplifier is neutralised by external feedback between the collector and base of the transistor. The degree of neutralisation required depends on the individual transitor, and provision has been made for this by the use of a small adjustable capacitor. With proper adjustment of the feedback capacitor, the radio-frequency stage is stable over the range from 90 to 130 kilocycles per second.

The crystal oscillator uses the same crystals as the *SR.24* receiver, but in the series-mode condition, and provision is made for adjusting each crystal to its nominal frequency.

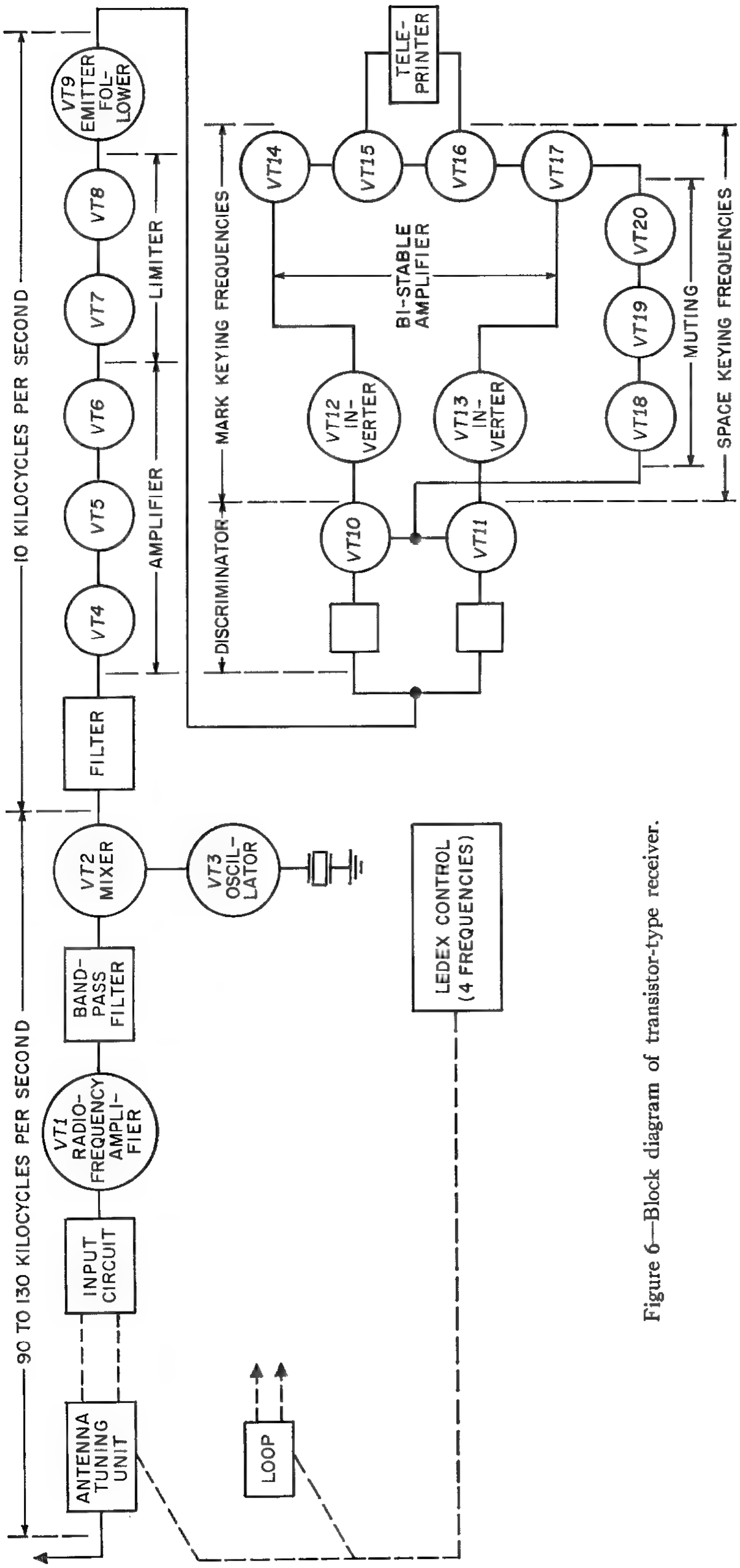


Figure 6—Block diagram of transistor-type receiver.

The mixer stage is coupled into a 5-element hermetically-sealed filter having a bandwidth of approximately 100 cycles per second at the 3-decibel-down points and 400 cycles per second at 60 decibels down with a mid-band frequency of 10 kilocycles.

All frequency-determining components of the discriminator are mounted in a hermetically sealed box. Output keying levels at the mark and space frequencies go to bi-stable amplifiers.

To prevent the printing of random symbols during adverse conditions, one of the output transistors is blocked by a voltage derived from the discriminator, via a threshold control. The all-up weight of the receiver unit is approximately 9 pounds (4 kilograms).

6. *Field Trials*

6.1 Ground-to-Air Over the North Atlantic

Field trials of the system were carried out over the North Atlantic route under the sponsorship of the International Air Transport Association and in co-operation with the British and Canadian Governments and the British Overseas Airways Corporation. One ground station was located at Galdenoch in Scotland and radiated 1·5 kilowatts on 121·6 kilocycles per second. This power was later increased to approximately 2·5 kilowatts. The other ground station was located in Canada, at Chatham, some 400 miles (644 kilometres) inland and radiated approximately 5 kilowatts on a frequency of 118·8 kilocycles per second. A frequency shift of 40 cycles per second was used for the tests reported below.

The limit of range was determined by the point at which the copy had deteriorated to 90 per cent. This point is quite sharply defined; as a further small decrease in signal-to-noise ratio leads to complete failure.

The normal test procedure calls for reception from the nearer station with the change being made at 30 degrees west longitude.

As the tests proceeded, some improvements were made to the receivers and to the antenna system at Galdenoch. The copy obtained at 30 degrees west longitude for later flights was considerably better than 90 per cent. In fact, 99·8-per-cent copy was obtained from both the British and Canadian stations at 30 degrees longitude. When the equipment was left on the frequency of the British station, 98-per-cent copy was obtained at 35 degrees, and 90-per-cent copy at 40 degrees. These figures refer to night ranges.

When operating in daylight the performance is somewhat reduced, but better than 90-per-cent copy is obtainable at 30 degrees. This corresponds to a field strength of 15 microvolts per metre, assuming an effective height of 15 centimetres for the open-wire antenna system.

It should be remembered that the receivers used in these tests were very carefully tuned, and some deterioration should be allowed for in practical service. On the other hand, the Canadian station was located some 400 miles (644 kilometres) inland, and the power of both stations was below the 10 kilowatts recommended by Doutre, which seems to be a realistic figure.

The main causes of mutilation are as follows:—

A. Flying through clouds charged with static. In general, the time lost does not exceed 10 minutes, although this figure depends on the height of the aircraft relative to the height at the top of the cloud.

B. Fading at points where the sky and ground waves are in antiphase and equal in amplitude. This effect does not usually last for more than a few minutes.

C. Intense noise interference from thunder storms. This generally lasts for a period not exceeding fifteen minutes, although the governing factors in *A* will apply. Static crashes from distant storms may cause an occasional mutilation.

On occasions, aircraft noise has been found to be a limiting factor if the installation is defective. With adequate radio-noise suppression applied to the electrical apparatus and with meticulous care in the bonding of interconnecting cables, particularly those connected to the antenna coupling unit, aircraft noise effects are eliminated.

6.2 Shore-to-Ship Trials

In February 1957, an *SR.24* receiver was installed aboard R.M.S. *Queen Elizabeth.* A vertical aerial about 80 feet (24 metres) in length was used, but the tests were somewhat impaired by the length of the cable connection between the antenna and the antenna coupling unit.

The transmitter at Galdenoch was received and 100-per-cent copy obtained at a range of 900 miles (1667 kilometres). This fell to about 90 per cent at 1500 miles (2778 kilometres).

The trials were confined to an outward and homeward passage, and periods of severe weather conditions were encountered. No interference was experienced from any of the transmitters aboard the ship except when using hand-keyed transmission on 143 kilocycles per second at the limiting range.

An *RV.14* has been installed in a weather ship by the Ministry of Civil Aviation and at approximately 20 degrees west longitude, satisfactory copy obtained when receiving Galdenoch.

An *SR.24* was operated successfully from transmissions at Galdenoch at the recent Brussels Universal and International Exhibition.

7. *Acknowledgments*

The authors express their thanks to Messrs. T. Brunt and N. G. V. Anslow of the British Overseas Airways Corporation for an analysis of the ground-to-air flight tests and to Mr. Milton Dishal of ITT Laboratories for helpful correspondence.

Cryptographic Telegraph Equipment Mi544*

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METHODS of converting ordinary-language or numerical text to enciphered messages or cryptograms are generally grouped into two classes. In the first class, the elements of the original message are jumbled beyond recognition; however, the security of messages so enciphered is low and none of the methods in this class is compatible with teleprinter operation. The other class comprises the various methods of substitution. Some methods of substitution offer limitless security; moreover, the sequential operation of most methods in this class is consistent with the sequential operations of teleprinting and of typewriting or writing in general. The *Mi544*, an enciphering and deciphering equipment for teleprinters, provides the ultimate degree of security. No additional time is lost as a result of the use of ciphers and the *Mi544* can be included in any teleprinter circuit using the 5-unit code.

Enciphering by substitution requires that the original message (plain-language text, numbers, abbreviations, commercial codes, et cetera) be converted to a code determined by the carefully guarded key or cipher. The result is the enciphered message. At the receiving end, the same cipher converts the enciphered message back into the original message.

For processing by machines, each element of both the original message and the cipher must have a numerical value of some kind. This condition is readily met by the teleprinter code where each element of a 5-unit character combination has the value of either 0 or 1. The fact that the binary system of numbers is employed greatly facilitates enciphering and deciphering and does not in the least affect the security of a ciphering system. Security depends solely on the fact that the cipher consists of a practically endless sequence of coded characters statistically distributed and appearing at irregular intervals.

The preparation of the cipher is an important detail of this ciphering method. There are ciphering equipments that produce the cipher simultaneously with ordinary telegraph operation. They are rather complicated and require operation and maintenance by highly skilled personnel.

1. *Philosophy of Design*

A few historical notes will show that the modern trend is to prepare the cipher separately and to use it, whenever required, in equipment like the *Mi544*.

During 1915 to 1925, methods and devices became known by which, apparently for the first time, ciphers were stored in the simplest form; that is, by perforations in a paper tape. This cipher was used simultaneously with the taped original message to transmit an enciphered version of the latter. Since a cipher of exactly the same form had to be available at the receiving end, the problem arose of producing duplicate perforated tapes. A collateral problem was the physical transfer of the duplicate tapes to the terminal stations without disclosing the ciphers to unauthorized persons.

Events of the years 1915 to 1918 favored the development in the United States of an equipment in which existing units of printing telegraph apparatus were used as building blocks. The result was a device operating on the start–stop principle with the 5-unit code. The original message was manually punched in a tape. The cipher was also punched in a tape. Both tapes were sensed, each by its own transmitting distributor, the 5 contacts of the one being connected with the 5 contacts of the other. This connection employed relay circuits ensuring that two current pulses from the two machines or two no-current intervals resulted in state A (+), state B (−) being established whenever the two tape elements sensed were of differing polarity. The principles of this operation (see Figure 1) are simple for both enciphering and deciphering;

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* Originally published under the title, "Das Mischgerät Mi544," in *SEG-Nachrichten*, volume 4, number 4, pages 181–185; 1956.

the circuits built into the *Mi544* are based on the same method.

Table 1 shows this scheme in the enciphering and deciphering of the 5-unit code.

Views on the cipher composition to prevent unauthorized deciphering have often changed. Originally, the cipher was relatively short; this involved repeated use of the same cipher at the expense of the inviolability of the method and the ciphered messages. The next step was to use two or more ciphers whose elements were combined by several automatic transmitting distributors. The process of combination was governed by partly regular, partly irregular movements; this and the arbitrary selection of starting points along the cipher tape greatly aggravated matters for the unauthorized intercepting decipherer. The problems of counting the steps along the cipher tape and of marking them by printed numbers were solved satisfactorily. However, expert cryptographers worked out more-and-more-stringent specifications, one of them being that a cipher should be destroyed immediately after use. Another is the purely random statistical distribution of cipher elements. A modern electronic cipher generator has been described.[1]

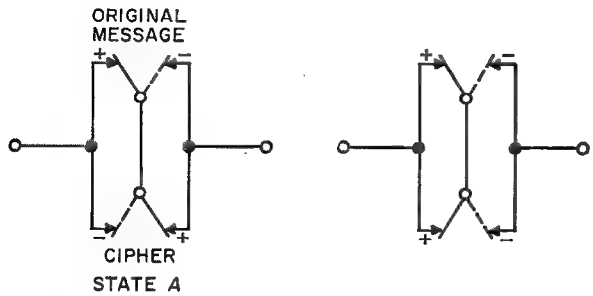

Figure 1—Basic enciphering–deciphering circuit.

TABLE 1

CIPHER PROCESS USING 5-UNIT CODE

Enciphering		Ciphered Message	Deciphering	
Original Message	Cipher		Cipher	Original Message
+	+	+	+	+
+	−	−	−	+
−	+	−	+	−
−	−	+	−	−

The above method of processing the cipher found wide application in the years 1930 to 1950. Several types of devices[2–10] operate on the indicated principles. These are generally receivers and transmitters combined with electric enciphering–deciphering features. Depending on application, the original message is prepared manually with tape transmitters or page printers or even with one or more manual perforators. The enciphered message can be directly transmitted to line or stored in the perforated tape for shipment to the recipient. The receiver can be switched to direct deciphering, producing the original message while receiving the enciphered message; or it may operate as a receiver, perforating a tape containing the enciphered message for deciphering later. These modes of operation are sketched in Figure 2.

Lately, consideration has again been given to the feasibility of combining all these functions in a ciphering teleprinter. In practice, the commercial teleprinter would then incorporate quite a few additional units, of which the most important are:

(A) A tape scanner for the original message.

(B) A tape scanner for the enciphered message.

(C) A storage register for the ciphering process.

(D) A tape transmitter.

(E) A reperforator.

(F) A built-in power supply for all functions.

[1] "Gerät zur Erzeugung von zufallsmässigverteilten Impulskombinationen für die Verschlüsselung von Fernschreibnachrichten ("Würfel-Locher")," *SEG-Nachrichten*, volume 4, number 4, pages 188–190; 1956.
[2] Austrian Patent 91 059; July 15, 1922.
[3] Austrian Patent 92 163; September 16, 1922.
[4] German Patent 355 393; June 6, 1920.
[5] German Patent 364 184; June 6, 1920.
[6] German Patent 452 194; March 21, 1926.
[7] United States Patent 1 516 880; November 18, 1924.
[8] United States Patent 1 522 775; January 13, 1925.
[9] G. E. Vernam, "Cipher Printing Telegraph System," *Electrical Engineering*, volume 45, pages 109–115; February, 1926.
[10] F. L. Rhodes and J. J. Carty, "50 Jahre Fernsprecher in den USA im Frieden und im Kriege," Verlag für Wissenschaft und Leben, Georg Heidecker, Berlin, Germany; 1934: see pages 145–146.

(G) Switches and monitors for the various modes of operation.

The assembly would be of substantial size requiring particular skill for operation and maintenance. An occasional trouble would affect the whole equipment, including the teleprinter functions. Many page printers now in the market would be useless as replacement units. Numerous special units now in use as, for instance, automatic transmitters, receiving perforators, manual perforators, et cetera, would and to varying degrees of mobility; units that are not required for a certain period need not be moved, operated, or maintained.

From the viewpoint of civilian application, it seems a technically and financially sound policy to retain existing teleprinters and to provide attachments for secret communication. For all these reasons, the *Mi544* was developed as an independent attachment. It operates on electromechanical principles using subassemblies and units that have for many years proved their value.

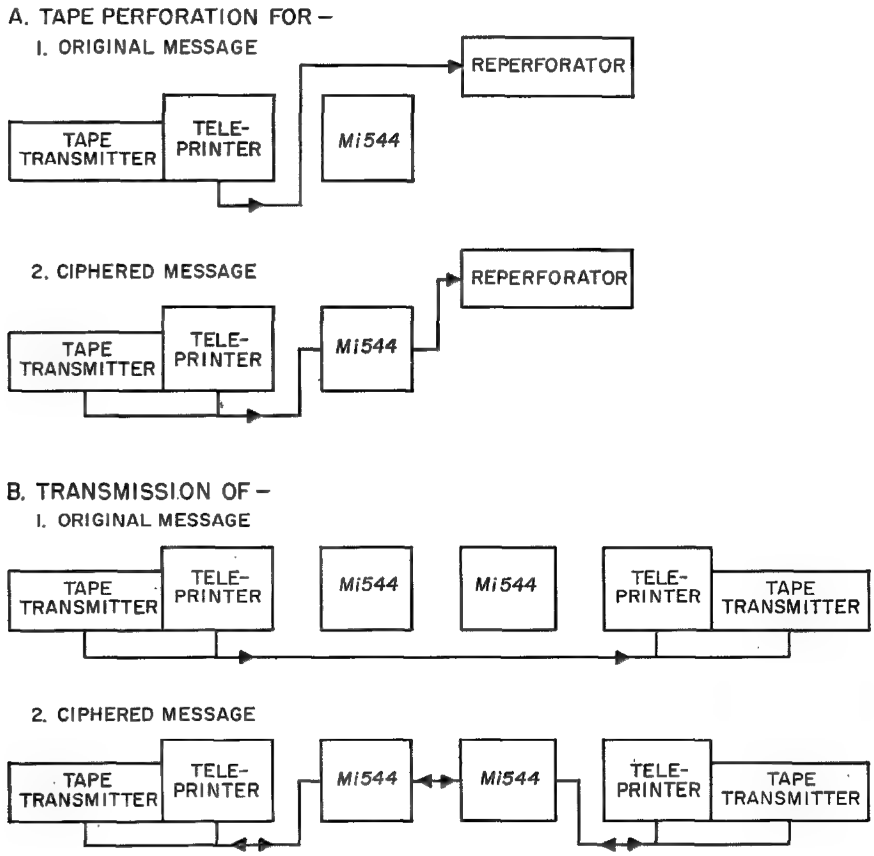

Figure 2—Teleprinter transmission of original and ciphered messages.

be of no practical value in such a mode of operation.

Things look quite different, however, if the encipherer–decipherer is constructed as a separate, selfcontained equipment. By electrically connecting this equipment to the existing input–output equipment, all the functions described and more can readily be exercised. Such a system is adaptable to varying operating conditions

2. Equipment Description

Figure 3 shows the *Mi544*. Below the control panel with its two switches and two indicator lamps, the unit sensing the cipher is mounted. The top has a recess in which the cipher tape roll is inserted. A metallic magazine encloses this roll and can be locked to prevent unauthorized removal of the roll. Special blocking means

preclude the possibility of rewinding the roll and using it a second time.

Figure 4 shows the equipment with roll and hood removed. The two receiving systems are immediately behind the indicator lamps and toward the rear, under the covers, are the driving motor and control relay group. Several receptacles for the connecting local teleprinters and an ammeter monitoring the telegraph circuits are mounted on the left-hand and front apron of the chassis.

Figure 3—The *Mi544* equipment.

The *Mi544* connects to 220-volt 50-cycle-per-second mains supply. The receptacles in the chassis permit connection of the local teleprinter, the transmitting distributor, and, if necessary, the receiving perforator. Plugs are used for connection to the transmission line or to the subscriber box (in the case of subscriber dialing) respectively.

2.1 Operational Sequence

When a connection is established, the red indica-

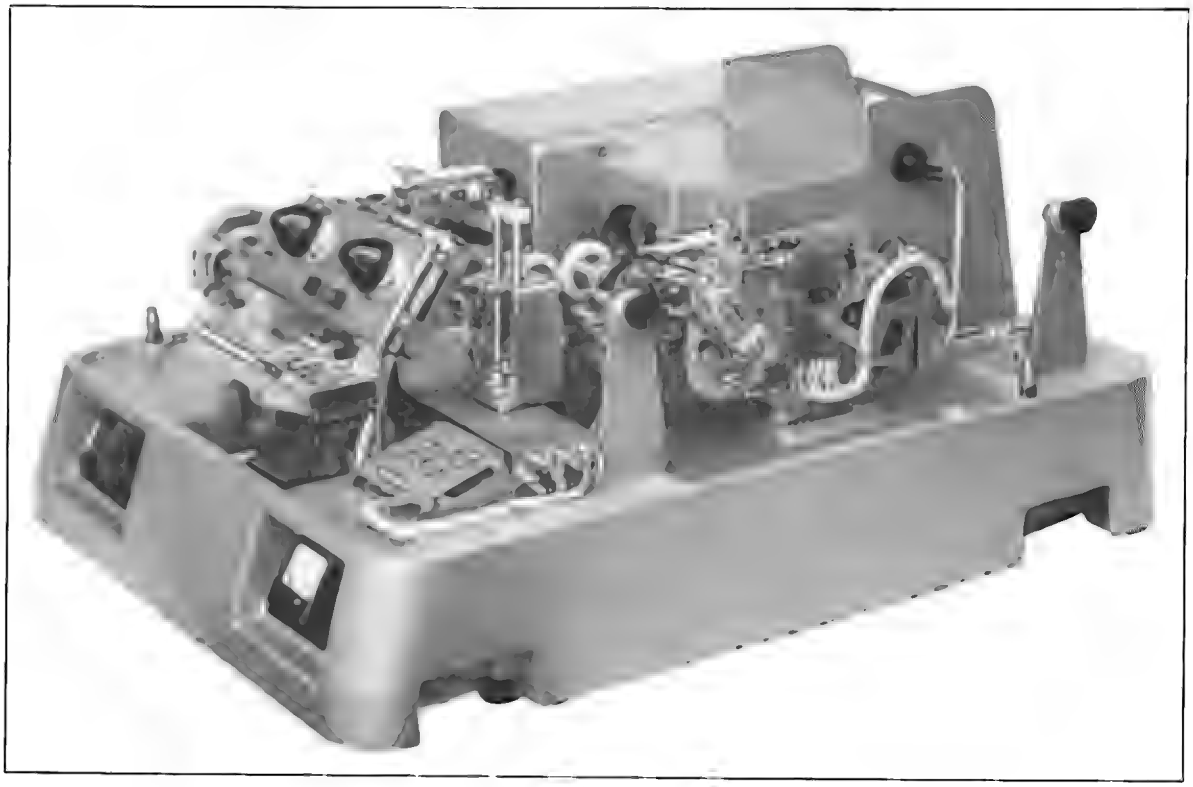

Figure 4—The *Mi544* machine with hood removed.

tor lamp signals and the teleprinter is started. The line is then connected to the teleprinter and the tape transmitter; the subscribers may now communicate in ordinary-language typewriting. When both subscribers have inserted their ciphers at the agreed-upon starting point, the ciphering devices of the *Mi544's* are switched on. The red lamps go out and the green lamps light; at the same time, the *Mi544* motors are started. If the first message character is now keyed on the teleprinter keyboard, the first character of the cipher is also sensed, combined with the teleprinter character of the original message, and transmitted as the enciphered character. At the end of the first character, the cipher tape is advanced to the next character, which is combined with the second character of the original message, and so on. At the receiving end, the cipher is sensed in the same way and used to decipher the incoming enciphered message. The original message character thus recovered is passed to the teleprinter for printing and the cipher tape is fed to the next character position.

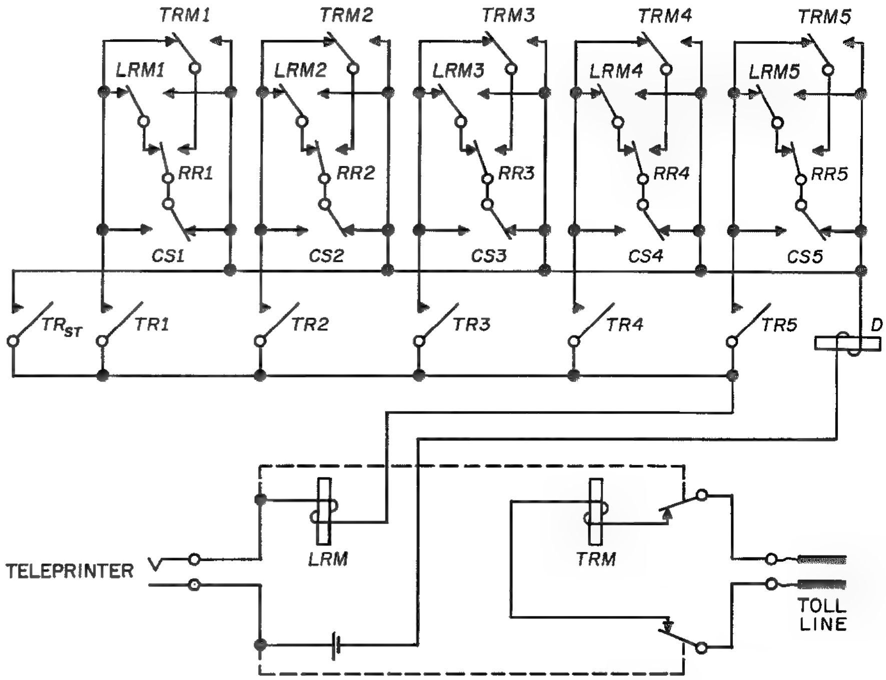

Figure 5—Basic circuit of *Mi544* connected for deciphering. *LRM1* to *LRM5* = storage contacts of the local reception magnet *LRM*. *TRM1* to *TRM5* = storage contacts of the toll reception magnet *TRM*. *CS1* to *CS5* = cipher-tape-sensing contacts. *TR1* to *TR5* = contacts of the built-in cam-controlled sender. TR_{st} = start–stop contact of the sender.

2.2 Circuit

Figure 5 shows the basic mode of operation for deciphering. All collateral functions are neglected in this presentation.

For transmitting plain text, see dashed lines, the local equipments (page printer and tape transmitter) are directly connected to the toll line or to the reperforator.

For cryptographic transmission, solid lines, the *Mi544* is divided into two separate circuits. The local teleprinter units are connected to the local-reception magnet *LRM* of the *Mi544* by a two-wire nonpolar circuit also supplying power.

The toll-reception magnet *TRM* of the *Mi544* is connected to the toll line, again a two-wire nonpolar powered line from the next exchange or the telegraph translator. Both reception magnets have additional monitoring contacts that determine whether the first (start) pulse is received from the near end or from the far end. In the first case, requiring enciphering, the contacts *LRM1* through *LRM5* are connected to the cipher-sensing contacts *CS1* through *CS5* in accordance with the operation previously outlined in Figure 1. The resulting enciphered message is applied through the transmitter contacts *TR1* through *TR5* to the toll line.

For deciphering in the incoming direction, the toll reception magnet *TRM* operates and contacts *TRM1* to *TRM5* store the pulses of the received enciphered message. Contacts *CS1* to *CS5* decipher the message by the above method.

The pulses of the original message thus recovered are sequentially sent from the *Mi544* to the local or near-end teleprinter for printing. In this operation, contacts *LRM1* to *LRM5* are disconnected by the reversing-relay contacts *RR1* to *RR5*. The blocking and switching relays for proper timing of these operations are not shown in Figure 5.

Relay *D* is provided so that the operator at the receiving teleprinter can stop the transmission whenever his equipment is not ready for reception or the message received is faulty. He does this by depressing any one key of his teleprinter keyboard (for instance, SPACE). As a result, relay *D* in the transmitting station will release and stop the transmission by the following process.

At the transmitter, the armature of relay *D* is mechanically pressed against the core in the intervals between the stop pulses. In undisturbed operation, the stop signal is always represented by a current pulse. If, however, signals arrive from the receiving end (utilizing the usual half-duplex circuit), there is a very-high probability that the stop pulse will be interrupted or suppressed after a few characters have been transmitted. This causes relay *D* to release. The transmission can only be resumed by the conventional manipulations. This interruption of transmission ensures that the original message is not, by some error, transmitted in plain writing.

Another means of preventing unintentional transmission of the original message is a paper-control lever that, together with the tape-sensing pins, is periodically applied to the perforated cipher tape. This lever remains inoperative as long as the tape moves properly. If the cipher tape runs out or the web between the perforations is damaged, however, this lever actuates relay *D* described above. The *Mi544* is thus switched off and the local circuit is opened so that no transmission is possible.

3. Characteristics

The technical data of the *Mi544* can be summarized as follows:

Dimensions = 400 by 540 by 260 millimeters ($15\frac{3}{4}$ by $21\frac{1}{4}$ by $10\frac{1}{4}$ inches).

Weight = 30 kilograms (66 pounds).

Mains input = 220 volts, 50 cycles per second, 250 watts.

Motor = universal collector type.

Built-in power supplies = 60 volts at 0.5 ampere for switching circuits and 120 volts at 0.15 ampere for teleprinter circuits.

Telegraph current (local and far-end) = 0.04 ampere.

Telegraph speed = 400 characters per minute (50 bauds) or 368 characters per minute (45.5 bauds).

Maximum permissible receiving distortion = 40 percent.

It is possible to build in a device[11] for character-by-character operation as well as for synchronous operation.

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[11] W. Schiebeler, "Synchronzusatz zum Michgerät Mi544," *SEG-Nachrichten*, volume 4, number 4, pages 185–188; 1956.

United States Patents Issued to International Telephone and Telegraph System; February 1—April 30, 1958

BETWEEN February 1 and April 30, 1958, the United States Patent Office issued 57 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

P. R. R. Aigrain, Laboratoire Central de Télécommunications (Paris), Impulse Multiplying Arrangements for Electronic Computing Machines, 2 822 131.

M. Arditi, Federal Telecommunication Laboratories, Radio-Frequency Transducer, 2 825 875.

A. J. Baracket, Federal Telecommunication Laboratories, Montage Amplifier, 2 825 755.

A. J. Baracket and S. A. De Mars, Federal Telecommunication Laboratories, Light-Energy-to-Video-Signal Transducing System, 2 829 199.

A. H. W. Beck, Standard Telephones and Cables, Limited (London), Electron Discharge Devices, 2 829 299.

A. H. W. Beck, A. B. Cutting, and J. W. Rayner, Standard Telephones and Cables, Limited (London), Grids for Electric Discharge Devices, 2 825 839.

A. E. Brewster, Standard Telecommunication Laboratories, Limited (London), Electric Trigger Circuits, 2 832 899.

J. H. Bryant, Federal Telecommunication Laboratories, Traveling-Wave Electron Discharge Devices, 2 822 500.

J. H. Bryant, Federal Telecommunication Laboratories, Traveling-Wave-Tube Oscillators, 2 829 252.

P. F. C. Burke, Standard Telephones and Cables, Limited (London), Electron Discharge Devices, 2 822 492.

V. D. Carver and M. Liao, Farnsworth Electronics Company, Cabinet for Electronic Equipment, 2 823 973.

A. M. Casabona, Federal Telecommunication Laboratories, High-Frequency Hybrid Circuit, 2 822 525.

K. W. Cattermole, Standard Telephones and Cables, Limited (London), Electric Pulse-Time Modulators, 2 822 520.

R. F. Chapman, Federal Telecommunication Laboratories, Display Arrangement for Direction Finders, 2 825 901.

A. R. Denz, Kellogg Switchboard and Supply Company, Temperature-Compensated Direct-Current Transistor Amplifier, 2 830 257.

P. T. Farnsworth, Farnsworth Electronics Company, Storage-Type Electron-Tube System, 2 820 111.

G. C. Florio, Federal Telephone and Radio Company, Mold for Semiconductor Ingots, 2 825 549.

S. G. Fong and H. W. G. Salinger, Capehart–Farnsworth Company, Electron Multiplier, 2 824 253.

H. Grayson, R. A. G. Dunkley, and T. H. Walker, Standard Telecommunication Laboratories, Limited (London), Saturable-Core Transformer, 2 831 157.

S. J. Harris, Capehart–Farnsworth Company, Light-Distribution Measuring System, 2 831 054.

C. Heck, Süddeutsche Apparatefabrik (Nürnberg), Method of Producing Highly Permeable Dust Cores, 2 825 095.

A. Hemel, Kellogg Switchboard and Supply Company, Relayless Line Circuit, 2 828 365.

R. W. Hughes and R. L. Plouffe, Jr., Federal Telecommunication Laboratories, Pulse Generator, 2 829 346.

R. W. Hughes, R. L. Plouffe, Jr., and H. T. Peretko, Federal Telecommunication Laboratories, Pulse Generator, 2 829 282.

H. P. Iskenderian, Federal Telecommunication Laboratories, Traveling-Wave Electron Discharge Devices, 2 825 840.

J. Kalish, Federal Telecommunication Laboratories, Oscillator, 2 829 256.

A. G. Kandoian, Federal Telecommunication Laboratories, Artificial Load for Broad Frequency Band, 2 825 874.

W. Klein and W. Friz, C. Lorenz A. G. (Stuttgart), Electron-Beam Focussing Device, 2 828 434.

J. A. Kostriza and P. Terranova, Federal Telecommunication Laboratories, Line-Above-Ground to Hollow-Waveguide Coupling, 2 829 348.

J. Kruithof, L. J. G. Nys, and J. L. J. Donceel, Bell Telephone Manufacturing Company (Antwerp), Electric Switch, 2 822 431.

A. Lauterer, C. Lorenz A. G. (Stuttgart), Interlocking Electromagnetic Relay Structures, 2 825 239.

E. J. Leonard, Kellogg Switchboard and Supply Company, Intermittent-Flow Condenser-Storage Timer, 2 830 235.

D. J. LeVine and R. J. Merke, Federal Telecommunication Laboratories, Radio-Frequency Transducers, 2 825 876.

W. Lewanda, Federal Telephone and Radio Company, High-Voltage Rectifier, 2 832 923.

M. Lilienstein and A. W. Murphy, Federal Telephone and Radio Company, Regulated Power-Supply System Using Transistors, 2 832 034.

F. T. Littell, Federal Telephone and Radio Company, Grid Network for Pulsed Oscillator, 2 822 521.

S. Metzger, Federal Telecommunication Laboratories, Signal Level Coder, 2 832 827.

A. J. Montachausse and D. Dautry, Compagnie Générale de Constructions Téléphoniques and Le Matériel Téléphonique (Paris), Electromagnetic Relay, 2 824 923.

R. K. Orthuber, Capehart–Farnsworth Company, Automatic Focus Adjuster, 2 831 057.

E. F. Poppe, Bell Telephone Manufacturing Company (Antwerp), Thermoconductive Screening Can for Electronic Tube, 2 829 870.

J. C. Price, Standard Telephones and Cables, Limited (London), Electric Pulse Modulators, 2 822 522.

A. J. Radcliffe, Jr. and A. R. Denz, Kellogg Switchboard and Supply Company, Condenser-Timed Delayed-Signal Repeater, 2 830 128.

D. S. Ridler, Standard Telecommunication Laboratories, Limited (London), Multiple Telegraph Signal Regenerators, 2 828 358.

D. S. Ridler, R. Grimmond, Standard Telecommunication Laboratories, Limited (London), Electrical Information Storage Equipment, 2 825 890.

D. C. Rogers and P. F. C. Burke, Standard Telephones and Cables, Limited (London), Traveling-Wave Tubes, 2 824 996.

G. T. Royden, American Cable & Radio Corporation, Half-Duplex Telegraph Repeater, 2 829 200.

R. D. Salmon and L. B. Salmon, Creed & Company, Limited (Croydon), Printing Telegraph Apparatus, 2 827 511.

P. C. Sandretto, Federal Telecommunication Laboratories, Meteorological Radar, 2 822 536.

K. Sass, Mix & Genest (Stuttgart), Telephone Exchange Circuit, 2 830 126.

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F. Schmidt, Mix & Genest (Stuttgart), Circuit Arrangement for Transmitting Signal over Telephone Lines, 2 831 061.

W. Sichak and E. P. Westbrook, Federal Telecommunication Laboratories, Lens Antenna System, 2 822 541.

V. J. Terry, D. A. Weir, and N. F. Fossey, Standard Telecommunication Laboratories, Limited (London), Start–Stop Telegraph Regenerators, 2 822 422.

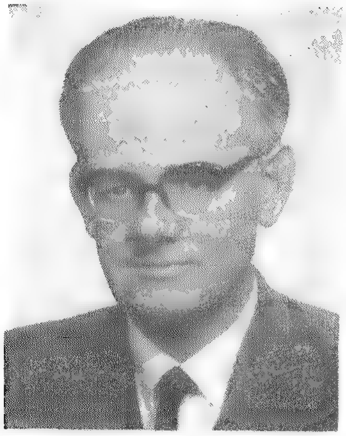
C. G. Treadwell, Standard Telephones and Cables, Limited (London), Electric Pulse Coding Arrangements, 2 825 873.

M. W. Wallace, Federal Telecommunication Laboratories, Cavity Magnetron, 2 828 444.

E. P. G. Wright, D. A. Weir, and J. Rice, Standard Telephones and Cables, Limited (London), Information-Handling Equipment, 2 831 179.

E. P. G. Wright, J. Rice, and R. C. Orford, Standard Telecommunication Laboratories, Limited (London), Electrical Information-Storage Circuits, 2 831 150.

Contributors in This Issue



Bruce Brooke-Wavell

Bruce Brooke-Wavell was born in Hove, England, on June 28, 1916. He studied mathematics and physics at Christ's College, Cambridge, and at London University where he obtained the degree of Bachelor of Science.

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Gerhard Grimsen

Mr. Brooke-Wavell contributes two papers to this issue—one describing the new model Seventy-five teleprinter, the other outlining the development of Creed telegraph apparatus during the past ten years.

• • •

Gerhard Grimsen was born in Braunschweig, Germany, on December 2, 1899. He studied at the technical college and university in Braunschweig and at the university in Berlin, from which he received a doctorate in 1922.

For four years, he worked in the Berlin central laboratories of the German Post administration.

Since 1926, Dr. Grimsen has been a member of the staff of Standard Elektrik Lorenz. He has worked on long-haul telephony and the construction of repeater offices, telegraphy, public telex networks, and ciphering systems. After the second world war, he was concerned with the reconstruction of the telex network and with special problems such as punched-tape engineering, enciphering and deciphering systems, and Comité Consultatif International Télégraphique et Téléphonique matters. He is author of the article on cryptographic equipment.

• • •

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On graduation, he joined Standard Telephones and Cables and worked on the original high-frequency transatlantic radiotelephone system. From 1929 to 1933, he was assigned to the Paris laboratories, where he was concerned with the Madrid–Buenos Aires high-frequency radiotelephone link.

He returned to Standard Telephones and Cables in 1933 and is now in charge of the design and development of point-to-point radio communication equipment.

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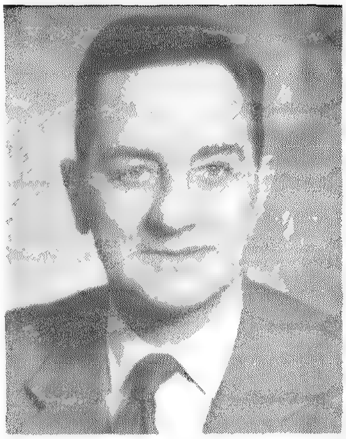
L. J. Heaton-Armstrong

• • •

John D. Holland was born in London in 1902. He was educated at the Salesian School at Battersea and then completed an apprenticeship at the Park Royal Engineering Works in 1922. He was awarded a National Certificate and a first-class pass in the final examination in radio communication of the City and Guilds Institute of London in 1939.

From 1922 to 1928, he served with the Royal Signals.



John D. Holland

He joined Standard Telephones and Cables in 1929 as an instrument maker and was transferred to development work in 1935. As a leader of a development group, he has been concerned with the design of receiving systems. Some of his recent work on receiving systems for amplitude- and frequency-modulated telegraph signals is treated in this issue.

Mr. Holland is an Associate Member of the Institution of Electrical Engineers.

• • •

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Dr. Kaiser is the author of the paper on the single-channel carrier telegraph equipment.

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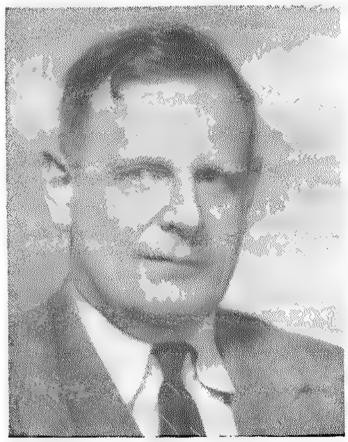

E. P. G. WRIGHT

E. P. G. WRIGHT joined the equipment engineering branch of the Western Electric Company in London in 1920, and in 1926 he took charge of project engineering when the company became associated with the International System. In 1928, Mr. Wright was transferred to the laboratories to work on switching problems.

He was in charge of switching system development for Standard Telephones and Cables Limited from 1932 until 1939. He was then appointed to the newly formed Standard Telecommunication Laboratories and has since been working on information-processing systems.

Mr. Wright is a Member of the Institution of Electrical Engineers and has three times been awarded the Fahie Premium by the Institution. He has been a delegate of the Comité Consultatif International Téléphonique for many years.

Mr. Wright is the author of two papers in this issue. One is on STRAD and the other on the calculation of storage requirements for electronic telegraph switching centers.

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Cryptographic Telegraph Equipment

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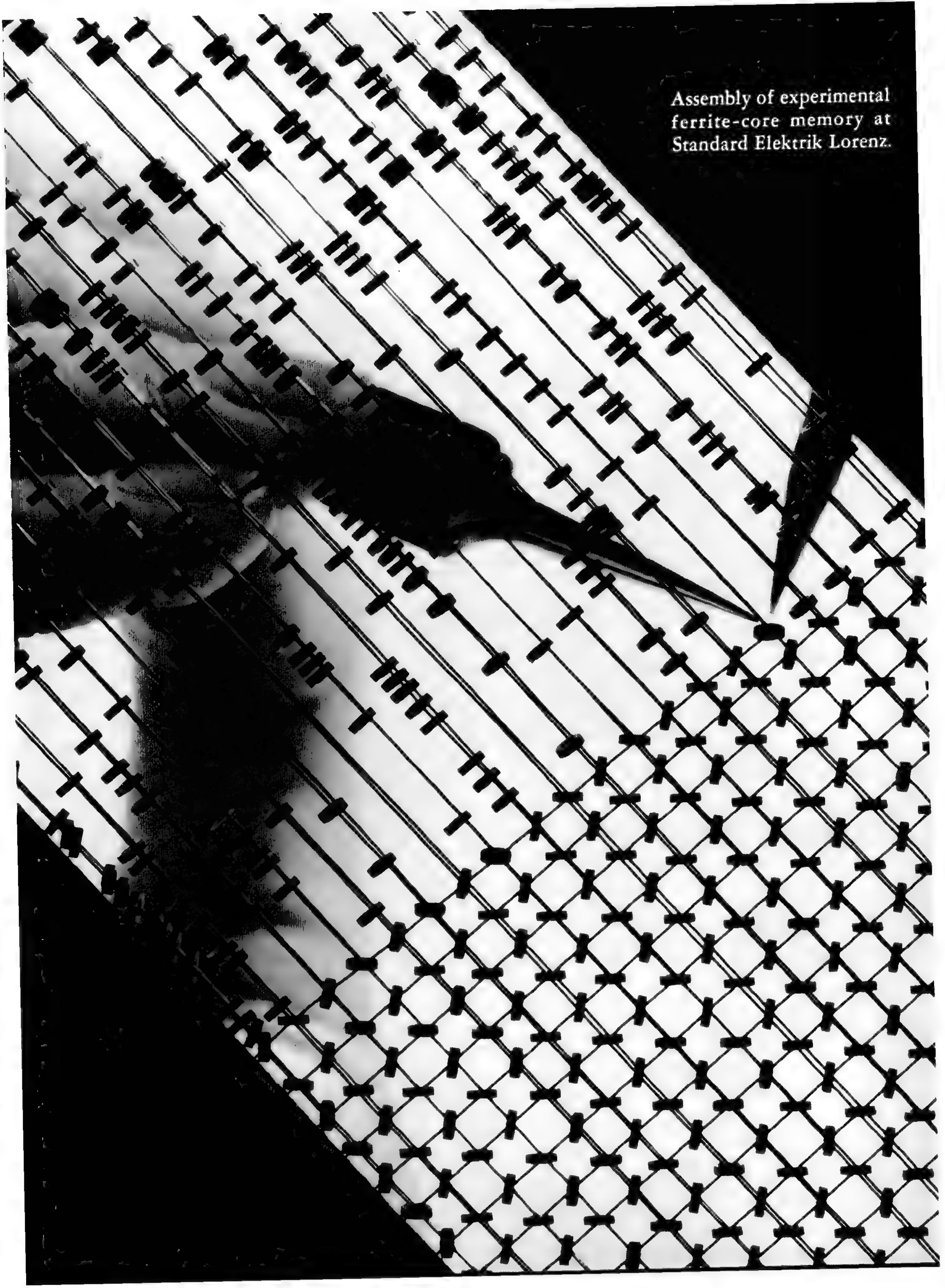


Assembly of experimental ferrite-core memory at Standard Elektrik Lorenz.

Coaxial-Cable Systems: Past and Future*

By A. W. MONTGOMERY

Standard Telephones and Cables Limited; London, England

SEEMINGLY, it is desirable to review some of the history of coaxial-cable systems if the trends of development that can be foreseen are to be correctly understood, since these spring from the developments of the past. Emphasis placed on certain characteristics of coaxial-cable systems has perhaps led to an unnecessarily restricted view of their possibilities. This augurs well for their future, since there remain many potential uses that invite examination.

A coaxial tube, it might be thought, is the obvious means by which telecommunication can be effected by cable. After all, a single wire in air with earth return has many of the features of a coaxial tube. Mathematically, all that is necessary to complete the tube is to wrap the earth around the wire at some suitable distance from it. This in fact was done to a reasonable approximation in an early use of coaxial cable for long-distance submarine telegraphy. In this, the sea, separated from the wire by an insulating layer, became what might today be called the return conductor.

1. *History*

The people who planned and laid the first transatlantic telegraph cable had tremendous faith, and when eventually this was justified by success, considerable study was made of the properties of coaxial tubes. This was in the middle and later 1800's. A considerable body of mathematical and other literature on the subject developed, of which the greater part is available and useful today.

Since the cable was a success for telegraph transmission, why then did it fail for speech? The answer, known to everyone today, is that speech requires a band of frequencies considerably wider than that needed for elementary telegraphy, and the attenuation of the cable rises very rapidly with frequency. It should be noted also that any bandwidth limitation is dependent on the rate of increase of attenuation with frequency, and not on any other essential characteristics of the theoretical design of the cable itself. This indicated that if amplification became available new possibilities would arise. However, before the invention of amplifying devices, it had been found that circuits could not be satisfactorily obtained from several wires in air along the same route and with earth return because of the resulting interference among them. Quite early, therefore, it became necessary to pair each wire with another to act as a return conductor.

As cable transmission became important, therefore, the same technique of using two wires per circuit naturally and correctly was adopted, and a very considerable art developed in the design of cables to attain smaller and smaller values of attenuation per unit distance and more and more freedom from interference among circuits.

From the theoretical work that had been going on, it was well known, however, that a return conductor completely surrounding a single wire and insulated from it would at high enough frequencies act as a sufficient shield to prevent interference between one tube and another. At this time (early 1920's) a cable system made in this way would have been entirely uneconomic and, as we know, an invention was awaited that would change the whole position. However, short lengths of coaxial cable were used for such purposes as carrying high-frequency currents to the antennae at the top of radio masts.

When the vacuum tube became readily available towards the end of the first World War, its usefulness was extremely great, but it suffered from the difficulty that its amplification–frequency characteristic showed considerable nonlinearity. It was eminently suitable for amplifying speech on existing types of cable pair, even though these required considerable attention to keep interference among circuits within bounds. And because the amplifiers could obviously amplify at frequencies considerably higher than those required by direct speech, attempts continued to be made to get more than one speech channel from each pair in a cable. (There had, of

* Based on a paper presented at the Christopher Columbus Celebrations in Genoa, Italy, on October 10, 1957.

course, been successful attempts before amplification became available to do this by means of such devices as phantom circuits, but these are no part of this theme.) The attempts had some success and attention was concentrated on packing channels as closely together as possible, consistent with satisfactory transmission characteristics, which by now could be specified with some exactitude.

The long-awaited invention that improved the linearity of amplification–frequency characteristics to an extraordinary degree, that is to say, the invention of negative feedback, immediately led engineers to consider means of getting more channels per pair, and so perhaps concentrated interest in only one of several promising directions. Attention also returned to the now-practicable use of coaxial tubes, which was attractive because of the potentially very large numbers of channels obtainable from them. It was apparent that a considerable amount of equipment would be necessary to locate an individual channel in its special place in the frequency spectrum. Terminal equipments therefore were from the beginning expected to be somewhat costly. The cost could be decreased by reducing to the minimum the amount of equipment required for each individual channel, and by making maximum use of equipment common to many channels.

These various concepts suggested that coaxial systems were most suited to routes over which large numbers of channels were required for long distances. The logic of this, however, obscured the more important truth that in fact the coaxial tube with its amplifiers was a means for obtaining a large bandwidth at a low cost, and that its economical use was not dependent on its use for large groups of channels and over long distances. What is clear, however, from the historical sketch given so far is how the earlier belief developed; although it should have been apparent even then that any coaxial-cable system has as its chief merit the provision of a wide frequency band at low cost.

2. *What is Bandwidth?*

It is desirable to remind ourselves of what we mean by a wide available band, and what are the causes that set any limit to the frequencies that can be used. The lower limit of frequency might be fixed by the point at which the loss of screening effect of the return conductor becomes important: but in fact some frequency higher than this (such as 60 kilocycles per second) is selected to ease the solution of equalisation problems. The higher limit may be set in practice by the effects of small departures from continuous cylindrical symmetry in manufacture or installation, which result in small impedance irregularities and consequent reflection effects. With careful design and manufacture, these can be kept under control to almost any desired extent; pulse testing methods give exact and extremely useful information both for design and inspection purposes. Figure 1 shows the results of a typical measurement. Although the variations from the mean value are very small, yet each can be assigned to a specific cause at a definite place in the cable should it be important to do so. Figure 2 shows an assembly of test equipment for the purpose.

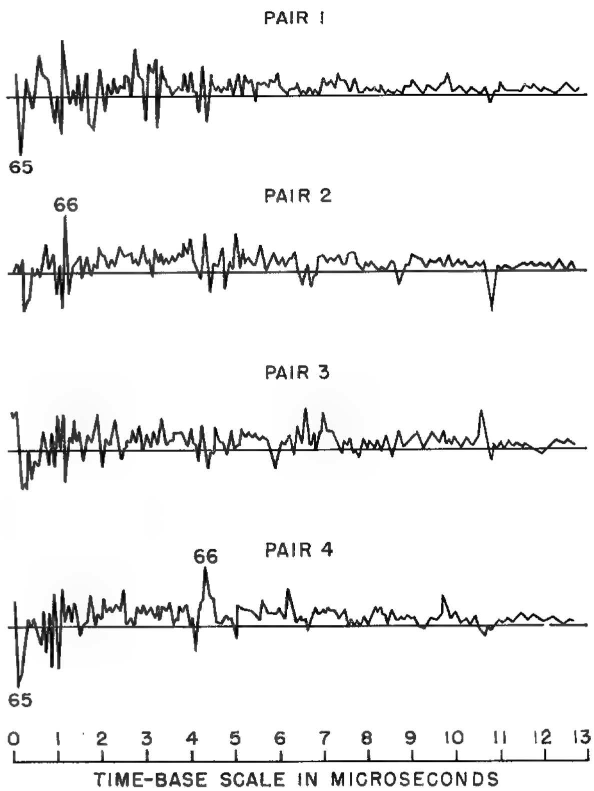


Figure 1—Oscillograms of reflections of a test pulse applied to an installed length of coaxial cable. The magnitudes of the largest reflections are stated in decibels below the incident pulse level. The incident pulse width was 0·1 microsecond.

In the end, however, the upper limit is fixed by economic considerations, compromise being effected by giving appropriate weight to factors such as required use of the circuits, possible designs of cable, spacing of repeaters, and design of the system as a whole to take care of known relationships between noise and amplification, and similar properties.

Again emphasising that a coaxial system is essentially an inexpensive one, nevertheless means were, and always will be, sought to reduce cost still further. A considerable saving was made by transmitting along the cable itself the power needed to operate at least some of the repeaters. Since the low frequencies were not in any case available for telecommunication transmission, this did not restrict the band. Accordingly the power losses that occurred along a cable of particular dimensions and the maximum voltage that could be impressed on the cable became factors in the design. Another series of factors was based on the maximum power output obtainable from a repeater, and in consequence the number of individual channels that could be obtained, making due allowance for the fact that all channels would not be in use at all times, nor would the powers in each channel be identical from moment to moment. There are many such sets of conflicting requirements, as there are in all telecommunication system designs, and these few are mentioned merely to illustrate the fact that with such a variety of factors to be taken into account undoubtedly many equally attractive solutions could be reached.

The solution by compromise is frequent in telecommunication problems and fortunately the chaotic conditions that could have resulted have been avoided by agreement among telephone administrations to recommend essential features for general adoption. The point to emphasise is that, through the medium of the Comité Consultatif International Télégraphique et Téléphonique and the wisdom of its members, it has been possible to select certain important features, possibly

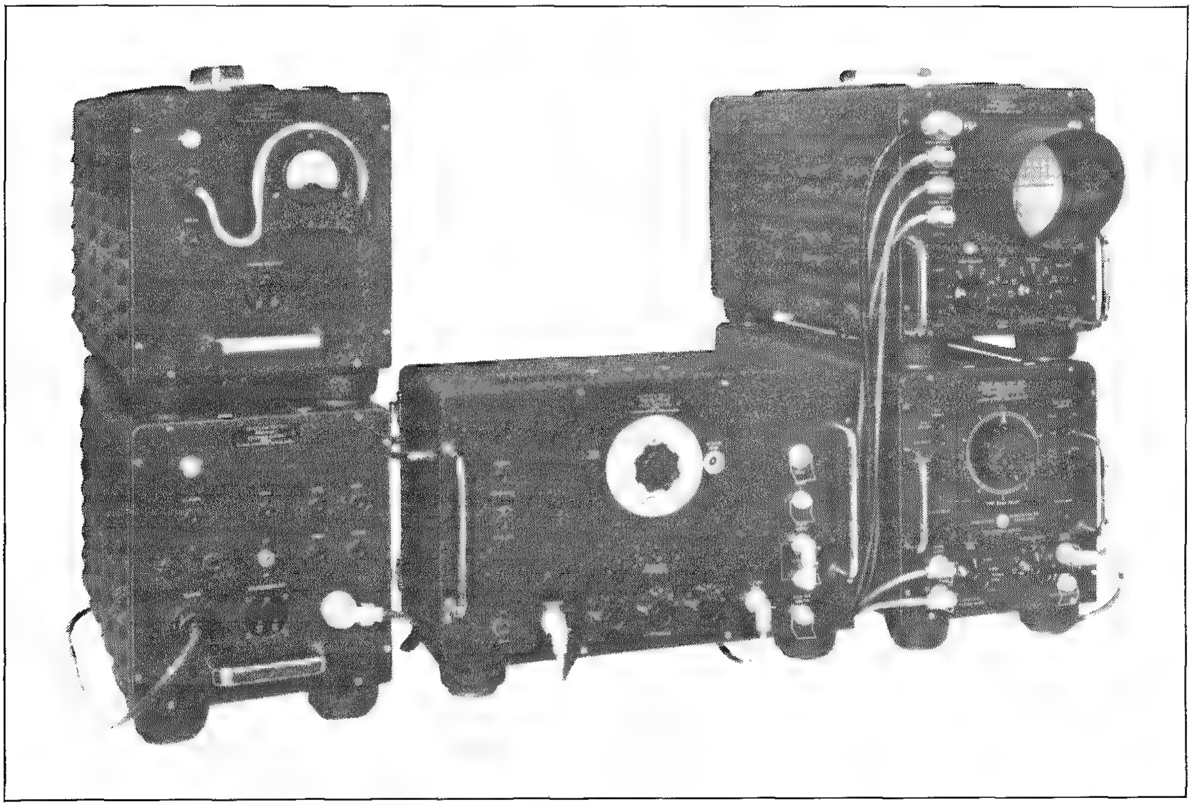


Figure 2—Assembly of equipment for making pulse tests on coaxial cable.

taken from a variety of types of system presented to them, to recommend these for general use, and so to enable world-wide communication to be effected most economically today and tomorrow. It obviously becomes increasingly important as invention makes communication more and more easy over greater and greater distances that requirements continue to be established wisely. They may not always be attainable immediately in their entirety, perhaps because of the cost of changing existing things, but they may eventually be met and maintained throughout the world.

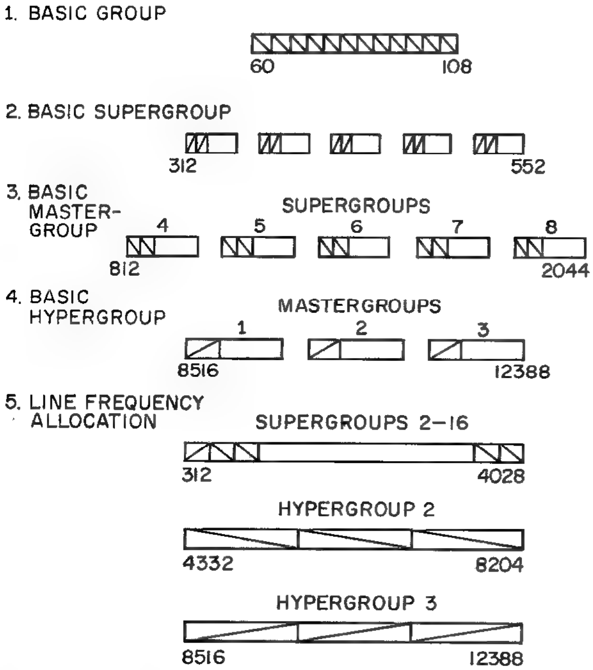


Figure 3—Wide-band coaxial telephone system adopted by Comité Consultatif International Télégraphique et Téléphonique. The line frequency allocation consists of hypergroups 2 and 3 and either hypergroup 1 or supergroups 2 through 16 as shown. Frequency limits are indicated in kilocycles per second.

3. *Increasing Frequency*

The extension of the transmission band into higher frequencies is already particularly important today, and not least in the field of coaxial-cable systems. For there are many ways in which progress is being made and bandwidth increased by one means or another. The continual increase in the usefulness of the bandwidth available from 0·375-inch (9·5-millimetre) coaxial cable has led to discussions by the Comité Consultatif International Télégraphique et Téléphonique concernin the use of frequencies up to about 12 megacycles per second for telephone and television systems, and provisional recommendations have been made. Figure 3 reminds us of the proposed frequency allocation, and illustrates well that it has been possible to build on the earlier recommendations as demands for channels have increased.

Equipment design becomes increasingly difficult as the frequency band increases, but remains within the bounds of economic achievement. Figures 4 and 5 show a typical amplifier already available to cover the 12-megacycle-per-second band. Despite the rapid growth of demands for channels, the 12-megacycle-per-second system should be adequate for some time to come. This is expected to be the case even when the transmission of colour-television signals is taken into account, since its bandwidth requirement is commonly expected to be within the limits of this system.

4. *Increasing Attenuation*

There is also another totally different approach to the quest for circuits. From the beginning, because the use of coaxial systems was thought of in terms of large groups of channels, there has been reluctance to part off from the cable small groups of channels at intermediate points. This could always have been done economically, but insistence on getting the maximum number of channels from a line has always led to some misgivings about this application. And as the demand for channels increases and pressure is brought to bear to obtain more and more channels from the available bandwidth, there develops even greater reluctance to attempt to drop small numbers of channels at intermediate points.

The question of the use of a coaxial-cable system for a relatively small number of channels becomes increasingly interesting. When considered in the past, as it was from time to time, the difficulty always proved to be that of providing an adequate number of repeaters of very small cost. Of course, the cost of a repeater must always include the cost of the building that houses it and

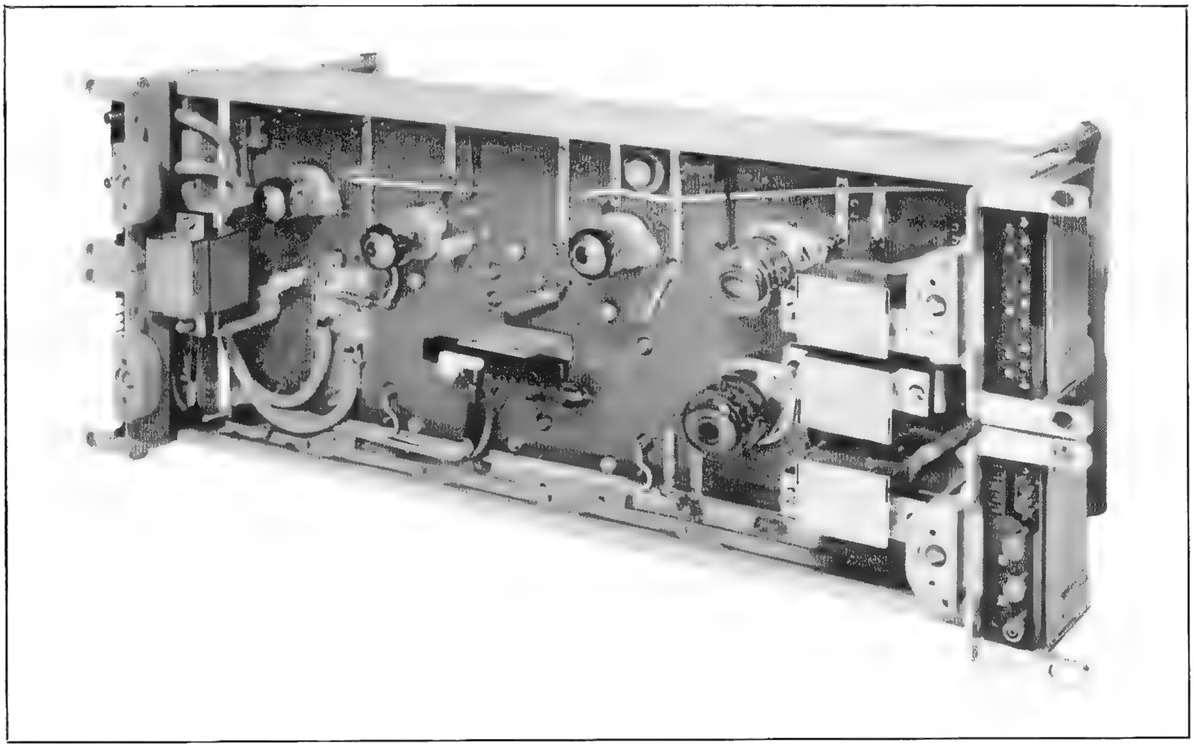


Figure 4—Amplifier for the 12-megacycle-per-second band that accommodates three hypergroups.

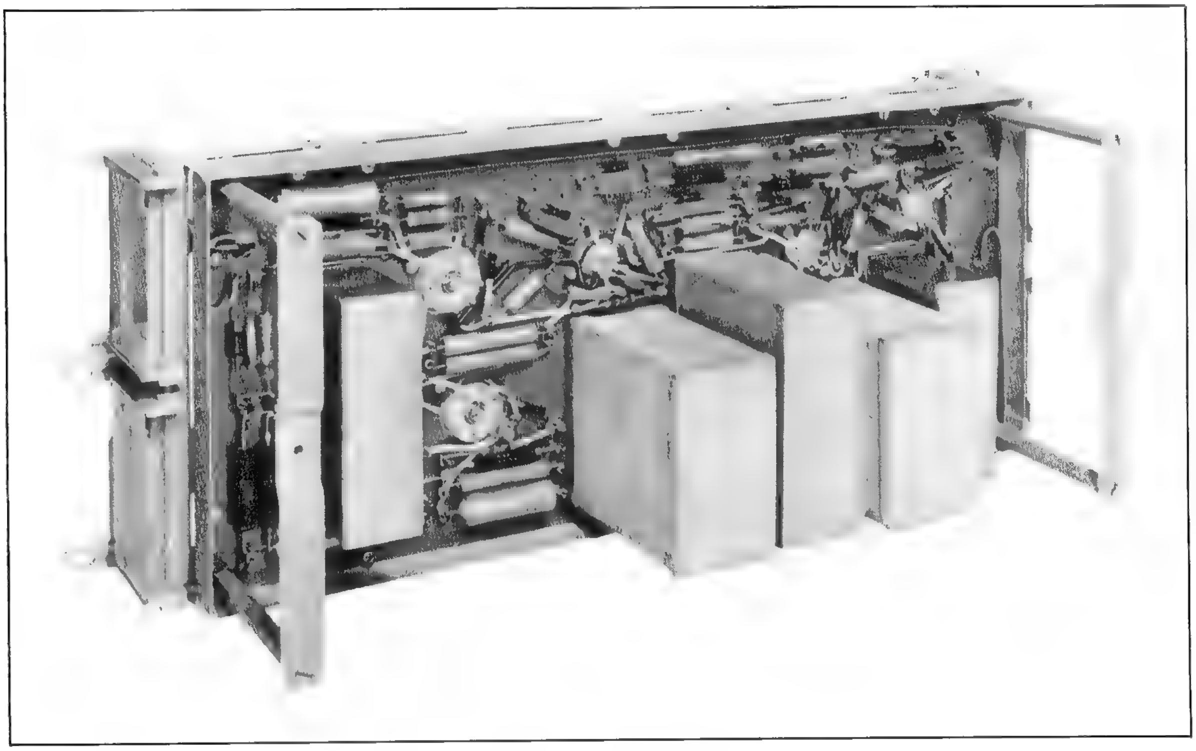


Figure 5 —View of reverse side of amplifier shown in Figure 4.

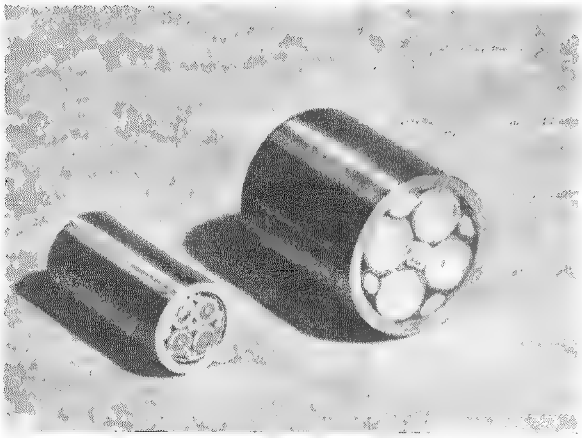


Figure 6—Cross-sectional view of 0·375- and 0·163-inch (9·5- and 4·1-millimetre) tubes in cables.

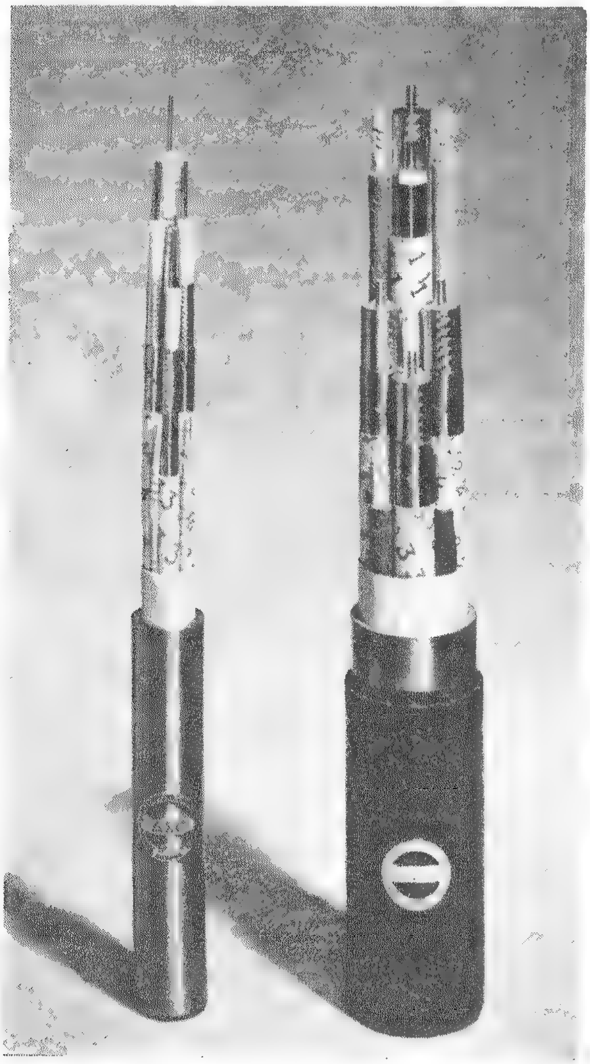


Figure 7 —Stripped view of the 0·375- and 0·163-inch (9·5- and 4·1-millimetre) cables.

of the power that must be supplied to it. Unattended repeaters, whose operating power is transmitted along the cable, have necessitated such total voltages as have limited reduction in size of the cable or, viewed differently, have imposed a limitation in the numbers of them that can be operated in tandem. An economic difficulty has always been that the 0·375-inch- (9·5-milli-

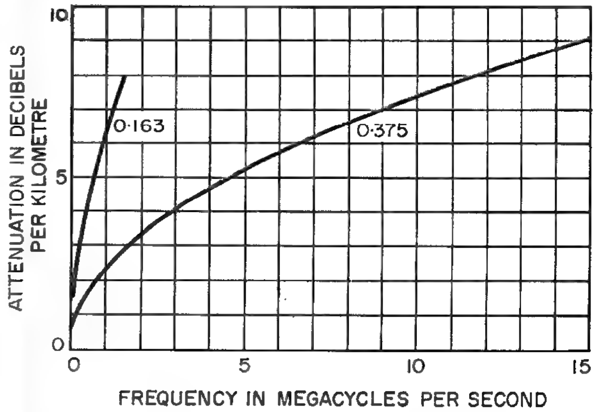


Figure 8—Comparison of the nominal attenuation at 10 degrees centigrade of the 0·163- and 0·375-inch (9·5- and 4·1-millimetre) coaxial tubes are a function of frequency.

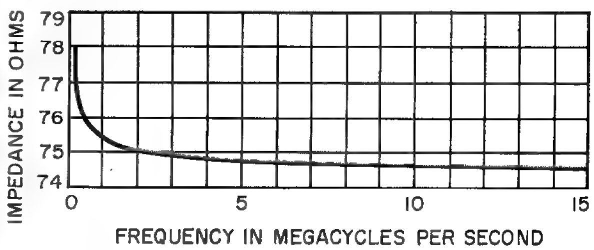


Figure 9—Nominal impedance of 0·375-inch (9·5-millimetre) tube versus frequency.

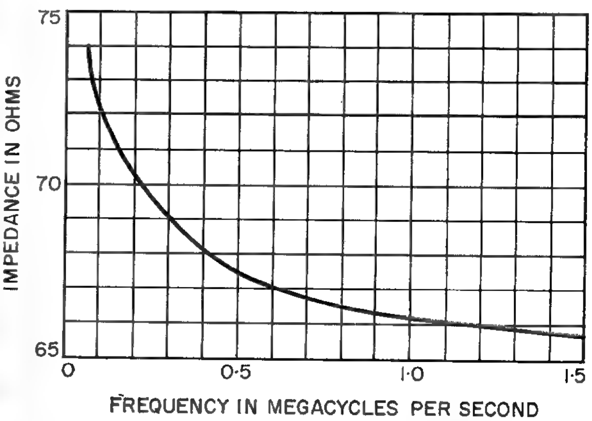


Figure 10—Nominal impedance of 0·163-inch (4·1-millimetre) tube plotted against frequency.

metre-) cable system is essentially economic even for quite short distances, and it has been correspondingly difficult to design a system that would show substantial economies if limited to even shorter distances.

The position changed considerably with the invention of the transistor, chiefly because it is essentially a low-voltage device. As the production of reliable transistors becomes assured, it is reasonable to contemplate their use in stable equipment. Because the transistor requires only moderate voltage and consumes little power two things change when comparison is made with valve-operated devices. These are the size of cable and the repeater spacing, both of which can be reduced without restricting in any way bandwidth availability. The function to be performed by these smaller cables would obviously be that for which the existing cables, however illogically, were considered not to be suitable. It would be reasonable to design initially for a limited number of circuits to be used over short distances. Cable attenuation could therefore be allowed to increase faster than the amplification obtainable by the insertion of additional repeaters.

From these considerations a system has been developed that is designed to provide up to 300 two-way speech channels, in general of Comité Consultatif International Télégraphique et Téléphonique quality, over a pair of coaxial tubes each of 0·163-inch (4·1-millimetre) diameter. The frequency band used lies between 60 and 1300 kilocycles per second and the repeater spacing has been fixed for this number of channels at approximately 12 000 feet (3·66 kilometres), this distance being a multiple of the accepted loading coil spacing of 6000 feet (1·83 kilometres). It is naturally not compulsory to use so many channels as this. Figures 6 and 7 show one version of the new 0·163-inch (4·1-millimetre) cable and a 0·375-inch (9.5-millimetre) cable for comparison. Figures 8, 9, and 10 show attenuation and impedance curves. Figures 11, 12, and 13 show the repeater and illustrate methods of connection and installation.

The transmission system itself is based on existing recommendations of the Comité Consultatif International Télégraphique et Téléphonique in that the basic 12-channel group in the frequency band from 60 to 108 kilocycles per second is used and supergroups are formed in the frequency band from 312 to 552 kilocycles per second in the

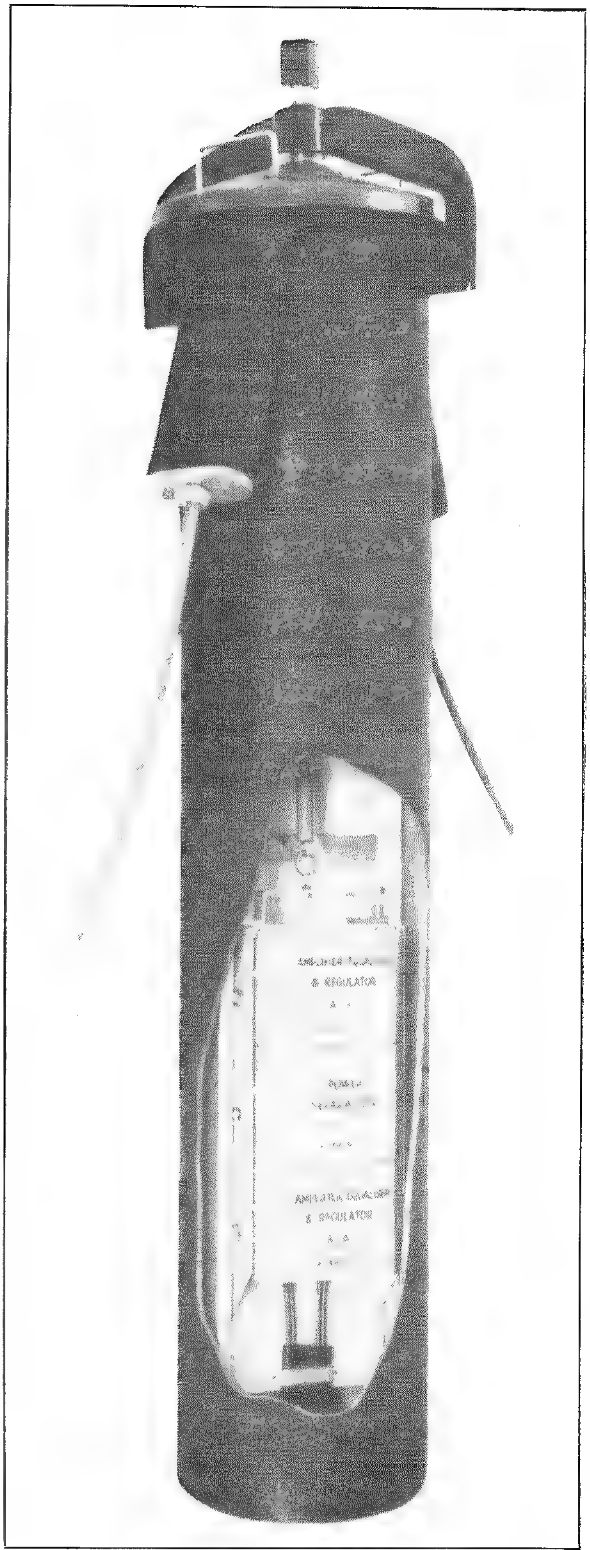


Figure 11—Two transistor repeaters and the power separation circuit for the repeaters are in a sealed housing.

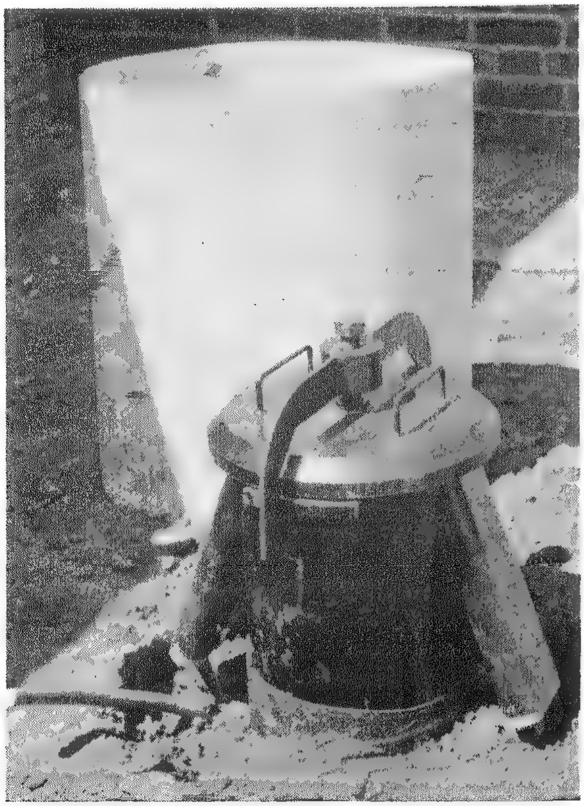


Figure 12—Repeater housing projecting from the ground.

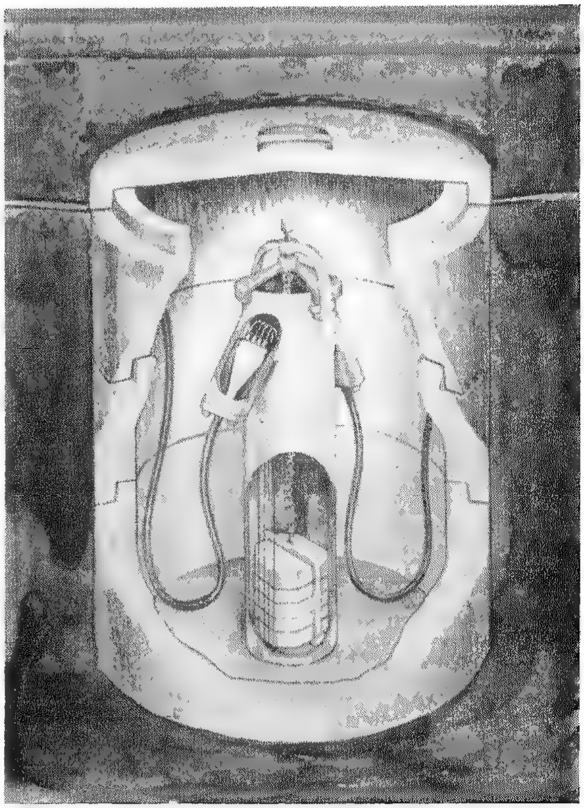


Figure 13—Cutaway view of repeater in manhole.

usual manner. Five of these supergroups may be used and will permit interconnection between this system and other approved systems by channel, group, or supergroup as desired. Equally, of course, programme channels can be obtained in a normal manner by combining 3 or 4 adjacent telephone channels.

5. *Future Requirements*

These two coaxial systems offer between them very great flexibility, which may well be very helpful in solving problems that arise with the increasing use of long-distance dialling by subscribers. Groups of channels of a variety of sizes and operating economically over a variety of distances become immediately available and accessible and can be rearranged very quickly to meet circuit requirements, whether caused by the growth of traffic or by the provision of additional exchanges of present or future types.

The future of coaxial-cable systems appears now to point in two main directions; *one*, in the use of higher and higher maximum frequencies for increasingly greater distances; and *two*, in the use of higher-attenuation cables providing wide frequency bands over shorter distances. Figure 14 is interesting in that it shows on the same graph the attenuation characteristics of the two cables mentioned, the frequency scales being different

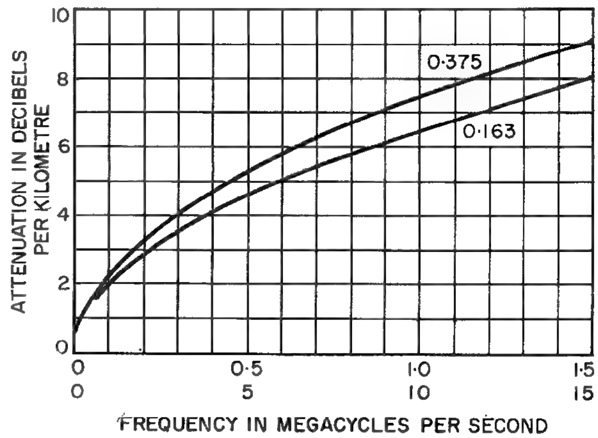


Figure 14—Nominal attenuation at 10 degrees centigrade for the two sizes of coaxial tubes. The higher-frequency scale is for the larger cable.

for the two curves. They illustrate once more that to speak of a wide band of frequencies is to speak of something that is adjustable as necessary.

However, reverting to the effects of history as outlined above, we may consider once more whether we are in fact not falling into our own trap by still regarding the coaxial-cable system as a means for providing a large number of channels closely packed in the frequency spectrum. Consequently, we should enquire whether it would not be interesting to look again at the economic use of the available wide bands. Because of the low voltage and power requirements of the transistor it may be reasonable to contemplate the use of repeaters at even closer intervals to make available even wider bands. Pulse systems, notably pulse-code-modulation systems, might then offer themselves as suitable for efficient economic use of the band, since one can often pay for simplicity of apparatus in bandwidth; and fortunately the transistor is eminently suitable for use in regenerative repeaters, which are reasonably simple. It remains to be seen however, whether the historical approach, which results in a demand for more and more channels from a given bandwidth, would allow a pulse-code modulation system to retain its simplicity, and correspondingly its economic justification.

6. Conclusions

From all of the above, one may conclude that the prospects for coaxial-cable systems are bright on account of the potentially available great bandwidths and the essential long- and short-term stability of the cable and its associated equipment. This means that whatever the type of transmission system evolved in the future or the demands made on any of them for increased maximum frequencies, there should be no difficulty in providing the appropriate transmission path. Also, the essential smoothness of the cable characteristics permits additions to equipment to be made that, at relatively small cost, will allow coaxial cables already installed to provide wider and wider frequency bands.

Perhaps the future prospects, too, are even brighter, since data and other information-transmission systems belonging to arts now in their infancy, will make increasing claim for very-stable circuits of as yet unknown bandwidth. Coaxial-cable systems with suitable maintenance are eminently adaptable to this type of requirement.

This paper has dealt only with main principles, which have been illustrated in part from existing practical embodiments of coaxial-cable systems. Having started with the deep-sea coaxial cable, it may be appropriate to terminate it in the same manner. As most excellently described previously by Sir Gordon Radley and Dr. M. J. Kelly, the use of amplification has already freed some more of the frequency band imprisoned in the submarine cable; and so the story feeds back to its beginning.

Some Modern Developments in Telegraph Transmission Equipment

BY W. F. S. CHITTLEBURGH

Standard Telephones and Cables Limited; London, England

EXPANSION of telegraph networks about 1930 was generally such that, because of the greatly increasing cost of telegraph operators, it became necessary to avoid manual retransmission. Administrations then had to decide between using some form of tape relay or introducing numerous separate point-to-point channels. Economic and other factors favoured the point-to-point arrangement, and at that time there was a universal changeover from time-sharing multiplex to frequency-division multiplex channels. The main trend in recent years has been towards longer point-to-point networks and the introduction of automatic switching of teleprinters. This has had two main effects on present-day voice-frequency telegraph design: firstly, a demand for the use of larger numbers of telegraph links in tandem, 6 to 8 in some cases; and secondly that the system be capable of resting for long periods in an idle condition, corresponding in amplitude modulation to a no-tone condition, and yet respond to the first incoming pulse with small distortion. It will be obvious that if 6 or 8 links are to be connected in tandem the distortion per link has to be kept low.

1. *Theoretical Considerations*

1.1 Telegraph Distortion

Telegraph distortion[1] can conveniently be divided into three components; characteristic, bias, and irregular or fortuitous distortion.

Methods of computing these various components into the probable distortion of several links in tandem are not yet fully established.[2] However both theory and experiment indicate that the characteristic distortion of n links in tandem tends to be n times that of one link, whilst irregular distortion is proportional to $n^{1/2}$. Though bias is algebraically additive it is due to random degrees of maladjustment, and the probable overall bias is again $n^{1/2}$ times the probable bias of one link.

The design of modern voice-frequency telegraph systems has been to keep the total distortion per link low, paying particular attention to the characteristic distortion, even to the extent of allowing interference from adjacent channels to increase somewhat; this involves the careful design of the amplitude–frequency and phase–frequency characteristics of the channels. The requirement of small distortion of the first mark signal after a long space condition has had its reaction on the design of the automatic level-adjusting means for amplitude-modulated systems.

1.2 Channel Filters

In achieving low characteristic distortion, a fundamental concept of channel filters has changed. A square-shaped discrimination characteristic, flat within the pass band, is no longer a neccessity. In fact it is the phase characteristic within the pass band that is more important; it should be kept reasonably linear. This concept applies equally to all forms of telegraph systems where the optimum ratio of signalling to bandwidth is required.

1.3 Amplitude Modulation

It is interesting to review some of the present methods that are, and can be, employed to convey telegraph signals through a channel using alternating currents restricted to a narrow band of frequencies.

Firstly with amplitude modulation the requirement that has arisen within recent years for the system to be capable of resting in the no-tone condition for long periods, and yet respond accurately to the first incoming pulse, has had its reaction on the automatic level-adjusting means. Most narrow-band systems relied, and

[1] Comité Consultatif International Télégraphique, "Draft List of Essential Telegraph Terms"; 1955.

[2] Comité Consultatif International Télégraphique et Téléphonique, Questions 8/9 and 9/9, Study Group 8; 1957–1960.

still do in some cases, on using a long time constant to "memorise" the incoming amplitude and to provide automatic gain and bias control. When tone was sent to line between messages it was only necessary to remember for the longest spacing period in a message, that is, 100 milliseconds or so. With very-long spacing periods this method of working became inadequate, and the next step was to put a delay in the signal path so that there was time to measure the incoming amplitude, set the automatic bias control, and then detect the signal. By this means quick-acting automatic level control was obtained that was capable of operating precisely on the first incoming signal after a long space.

Amplitude-modulated systems have been in use now for a sufficiently long period on both national and international networks that almost all the requirements[3] for such a system have been laid down. It is to be noted that although these requirements for amplitude-modulated equipment have changed little within the past few years, the actual performance of equipment being developed has improved to the extent that maintenance of these systems has been considerably eased.

1.4 Frequency-Shift Modulation

Though frequency-shift-modulated systems have been developed for several years now their use generally has been somewhat limited and the Comité Consultatif International Télégraphique et Téléphonique, taking advantage of these facts, is at present attempting to get agreement on the requirements for such systems when used internationally.[3,4] These systems possess the advantage of being very insensitive to changes of amplitude, but they are generally quite sensitive to frequency instability.

Frequency-shift modulation was for a long time considered only as a wide-band method, and only in relatively recent years has it been realised that it can give a ratio of working speed to bandwidth comparable to amplitude modulation. Narrow-band frequency-shift modulation, as used today, actually has an amplitude variation during the frequency transitions, and this is a vital part in the process for obtaining the full ratio of signalling speed to bandwidth.

It is well established now that for the same speed-to-bandwidth ratio a frequency-shift-modulated channel can have at least a 6-decibel signal-to-noise advantage over an amplitude-modulated channel. However, the fact that tone is always present in the frequency-shift-modulated case involves the penalty of increased power loading of the transmission media, and

Figure 1—View of Oban terminal station of the transatlantic telephone cable showing some of the frequency-shift-modulated voice-frequency telegraph equipments used between Great Britain and Canada.

[3] Comité Consultatif International Télégraphique et Téléphonique, Documents of VIII Plenary Assembly; 1956.

[4] Comité Consultatif International Télégraphique et Téléphonique, Question 16/9, Study Group 8; 1957–1960.

the 6-decibel advantage is reduced somewhat. The Comité Consultatif International Télégraphique et Téléphonique has agreed that for international working of a 24-channel system over a telephone circuit the level per channel of a frequency-shift-modulated telegraph system should be reduced by 1·5 decibels relative to the corresponding level of an amplitude-modulated system. Nevertheless it has also been agreed that on certain types of telephone circuits where the extra power loading is not important, for example, audio circuits and open-wire carrier circuits et cetera, the same level may be transmitted for frequency-shift-modulated as for amplitude-modulated channels, so that the full 6-decibel advantage may be maintained.[3]

Other advantages of frequency-shift modulation are that the level range over which the frequency-shift-modulated channels will operate is very wide, and that sudden changes of level cause only small increases of distortion, and these only when the change of level occurs at, or near, a signal instant. Unlike an amplitude-modulated channel, should the tone of a channel disappear, due to a fault condition, the output of the detector is left with no absolute control, and the channel becomes sensitive to noise. In this respect the Comité Consultatif International Télégraphique et Téléphonique has recommended that the equipment should be so designed that when such a fault occurs, the output of the receiver will be held in a start condition, and its response to noise prevented.[3]

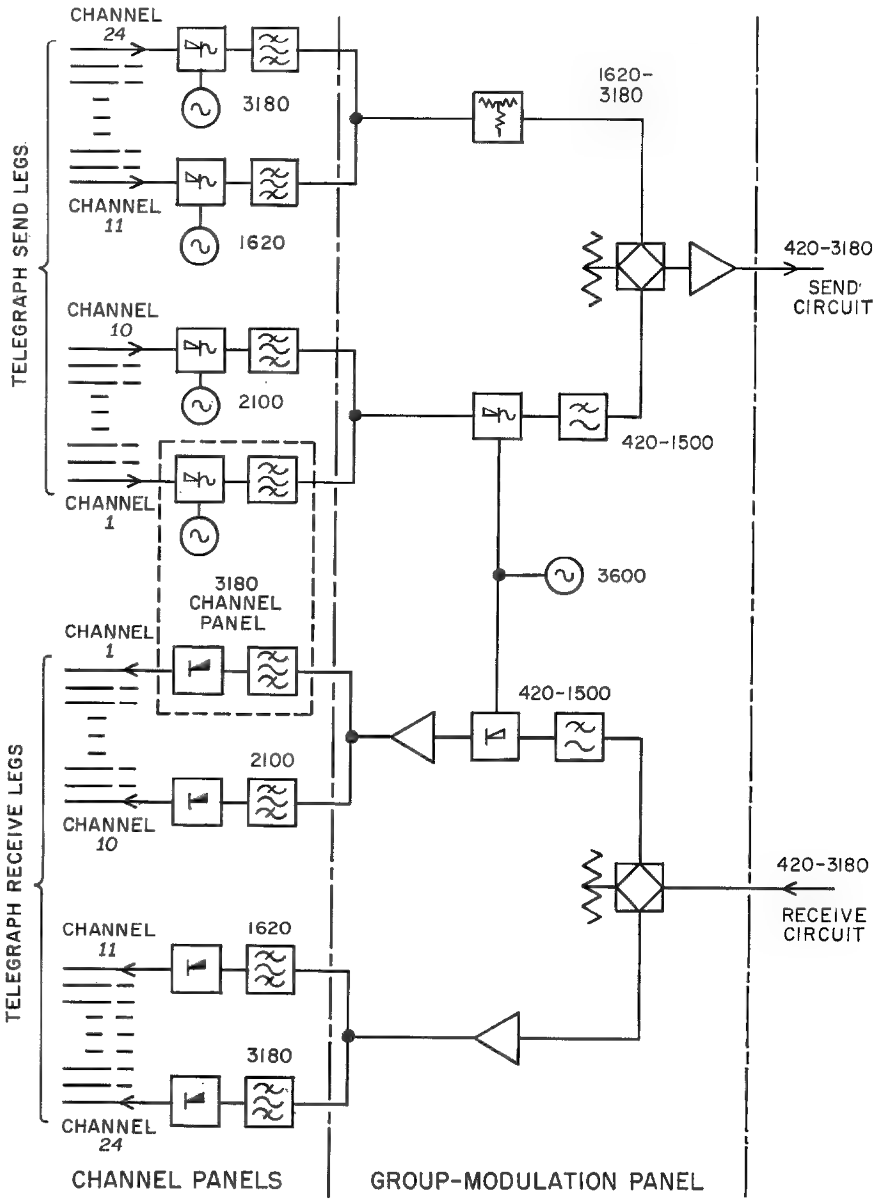


Figure 2—Group modulation arrangements in the *TA4* amplitude-modulated voice-frequency telegraph system. The frequency designations are in cycles per second.

1.5 Phase Modulation

Phase-modulated systems are extremely limited both in design and extent of application; as such they are, at the moment, of minor importance. These systems generally are of the phase-reversal type, the phase of a constant frequency being reversed at each transition from mark to space and vice versa. The disadvantage is that there is no reference condition to indicate whether a mark or a space is being sent. This means that should a transition be inserted or lost due to noise or other mischance the information is completely reversed from then onwards. Until now no solution, for start–stop operation, has been proved, though there is every hope that this may be done in the near future.[5]

There is a phase-modulated system[6] being marketed that is of interest. The individual voice-frequency channels are of a diplex nature, it being necessary to detect 4 different conditions of phase per channel. The system uses rhythmic transmission, and additional equipment is provided to convert, where necessary, the signals from arhythmic to rhythmic.

2. *System Features*

2.1 Sending Equipment for Amplitude-Modulated Systems

The older amplitude-modulated systems were designed, especially for large installations, to use multifrequency motor-generators as a source for the various channel tones, though in small installations individual oscillators for each channel could be provided. Though the multifrequency generator is still used, the tendency with modern systems is to provide individual oscillators for all channels, thus providing considerable flexibility, enabling small groups of channels to be assembled initially and to be increased as and when the need for more arises.

There is a technical reason also for the departure from multifrequency generators, especially from those that produce only the frequencies for channels 1 to 18. There is always a tendency when the lowest-frequency channels are directly modulated for the telegraph distortion of these channels to be somewhat greater than that on the higher-frequency channels; this is due to harmonic components of the modulating waveform interacting with the carrier frequency of a channel; it is known as carrier beat. There are methods whereby this interaction can largely be eliminated; in the *TA2* system, tones were modulated directly by means of a carefully designed double-balanced modulator, the direct-current input waveform on the receive leg being slightly filtered to remove the high-order frequency components. In the *TA4* system a simple modulator is used but only the tones of channels 11 to 24 are produced by direct modulation. Channels 1 to 10 are produced by group modulation of a second group of frequencies corresponding to channels 24 to 15. As will be seen in Figure 2, a similar group-demodulation process is performed at the receiving equipment. An advantage gained by this group modulation is that the coils and capacitors required in the channel equipment have convenient values and can be of small physical dimensions, enabling all channels to be made of the same minimum size.

2.2 Receiving Equipment for Amplitude-Modulated Systems

The design of the detector circuit of the early equipment has been described previously,[7] but it may be of interest to indicate briefly the circuits used for *TA2* and *TA4* equipments. In both of these the alternating-current amplifiers are linear over the operating range and are of constant gain; automatic bias control is used to compensate for changes of input level. The *TA2* equipment shown in Figure 3 utilizes a bias circuit having a quick charge and a very-long discharge time. The signal path includes a delay network, enabling the bias circuit to be charged to its correct voltage by the time the signal is applied to the output. This long discharge time on the bias circuit, though enabling the equipment to be designed and operated with very low characteristic distortion, is relatively inefficient when the mean incoming level suddenly drops;

[5] H. T. Prior et al, British Patent 693 704; September 8, 1950.

[6] E. T. Heald and R. G. Clabaugh, "Predicted Wave-Signalling Phase-Shift Telegraph System," *Communications and Electronics*, number 31, pages 316–319; July 1957.

[7] J. A. H. Lloyd, W. N. Roseway, V. J. Terry, and A. W. Montgomery, "New Voice Frequency Telegraph System," *Electrical Communication*, volume 10, pages 184–199; April, 1932.

the output signals suffer from a high degree of distortion until the bias circuit has discharged to its new correct voltage. Under normal conditions this system gives exceedingly good performance; its characteristic distortion for mixed messages is only about 3 per cent for 50-baud signals and 8 per cent for 80-baud signals.

In an endeavour to avoid the long time constant and at the same time to effect considerable simplification the *TA4* circuit[8] was developed. In this, as shown in Figure 4, the biasing circuit has two biases derived from the received signal; one for biasing the space–mark transitions, and the other for biasing the mark–space transitions. The circuit is so designed that, of the two biases, only that voltage that is the greater will be effective, each bias being independent of the other. The discharge time is made just sufficient to maintain the correct bias voltage for the instant of a mark–space transition, but not so long that it interferes with the new bias for the next space–mark transition. Thus every signal transition derives an individual bias, and very nearly instantaneous response of the equipment to sudden changes of level is achieved. Naturally if a change of level occurs at or near a signal instant, the modulation products of the level change will interact with those of the normal signal transition and may produce distortion, but this is liable to occur with any form of modulation when operating over narrow-band channels. The *TA4* system, because of its particular detector circuit, has been designed primarily

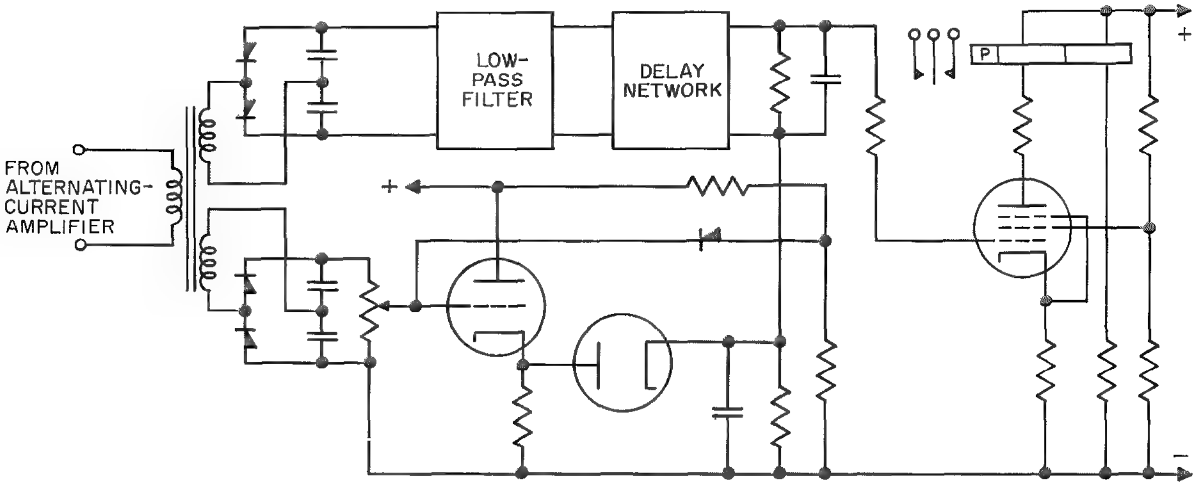
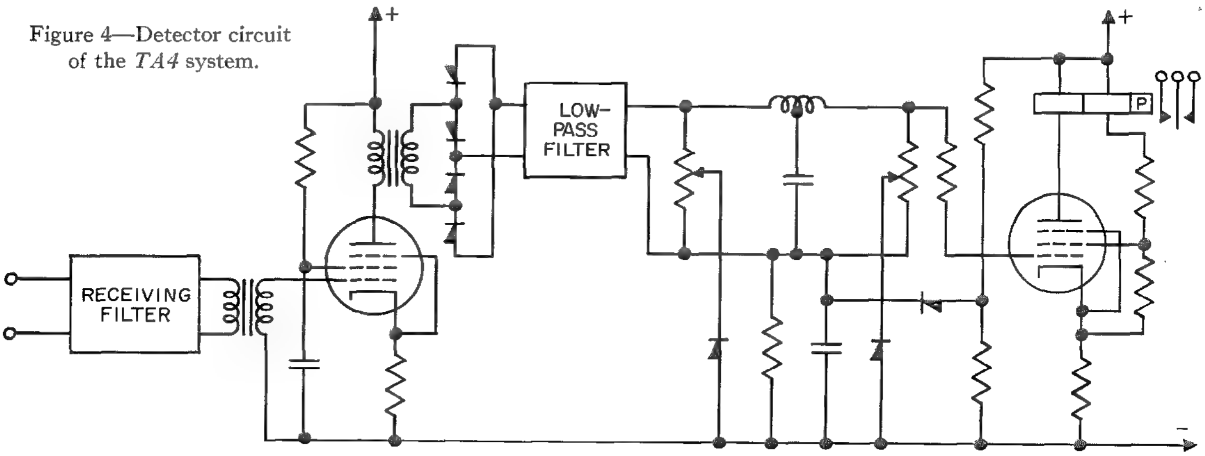


Figure 3—Biasing circuit of *TA2* amplitude-modulation detector.

Figure 4—Detector circuit of the *TA4* system.



[8] H. T. Prior, British Patent 693 769; September, 1950.

for 50 bauds, or lower, speed of working, especially if several such links are connected in tandem; but where only one link is contemplated higher speed may be employed. The characteristic distortion of this system at 50 bauds is about 3 per cent, and even with signals predistorted by amounts up to 35 percent the increase of distortion will not be more than the above characteristic distortion value.

2.3 Frequency-Shift-Modulated Systems

When the frequency-shift-modulated system *TF1* was developed[9] it was thought that its major application would be for extremely noisy line circuits or for radio working. To a large extent because of the proposed radio application, the channel equipment was engineered as two separate panels as may be seen at the left of Figure 5. The upper panel mounts the oscillator and modulator; the larger, the detector. The use of frequency-shift-modulated systems is now becoming more widespread for line working whilst for radio working there are further requirements, so the latest frequency-shift-modulated channels have been engineered with both send and receive circuits on the same panel.

To compensate for effects of variation of the ambient temperature, the *TF1* system has individually adjusted compensation circuits. It was arranged that the mean frequency of the oscillator in the modulator remain constant with change of temperature; but the compensation in the detector circuit did not attempt to correct the tuning of the discriminator coils, compensating instead the bias distortion caused by any drift of tuning due to temperature. Another refinement was the provision of a 300-cycle pilot channel to correct for any frequency drift that might have occurred between the modulation and demodulation carrier frequencies of the vehicle telephone channel. A frequency error from this source causes equal bias to the signals of all channels, thus, by detecting the drift over the pilot channel, a common compensating bias was applied to all channels.

The advent of improved materials for resonant units has made it possible to provide circuits for the *TF3* system that do not require special temperature compensation. Pilot equipment again can be provided, in this case at either 300 or 3300 cycles. As with the amplitude-modulated equipment *TA4*, to enable all channels to be of the same small physical size, channels 1 to 6 are derived by group modulation. Channels 19 to

[9] W. F. S. Chittleburgh, D. Green, and A. W. Heywood, "Frequency-Modulated Voice-Frequency Telegraph System," *Post Office Electrical Engineers Journal*, volume 50, pages 69–75; July, 1957.

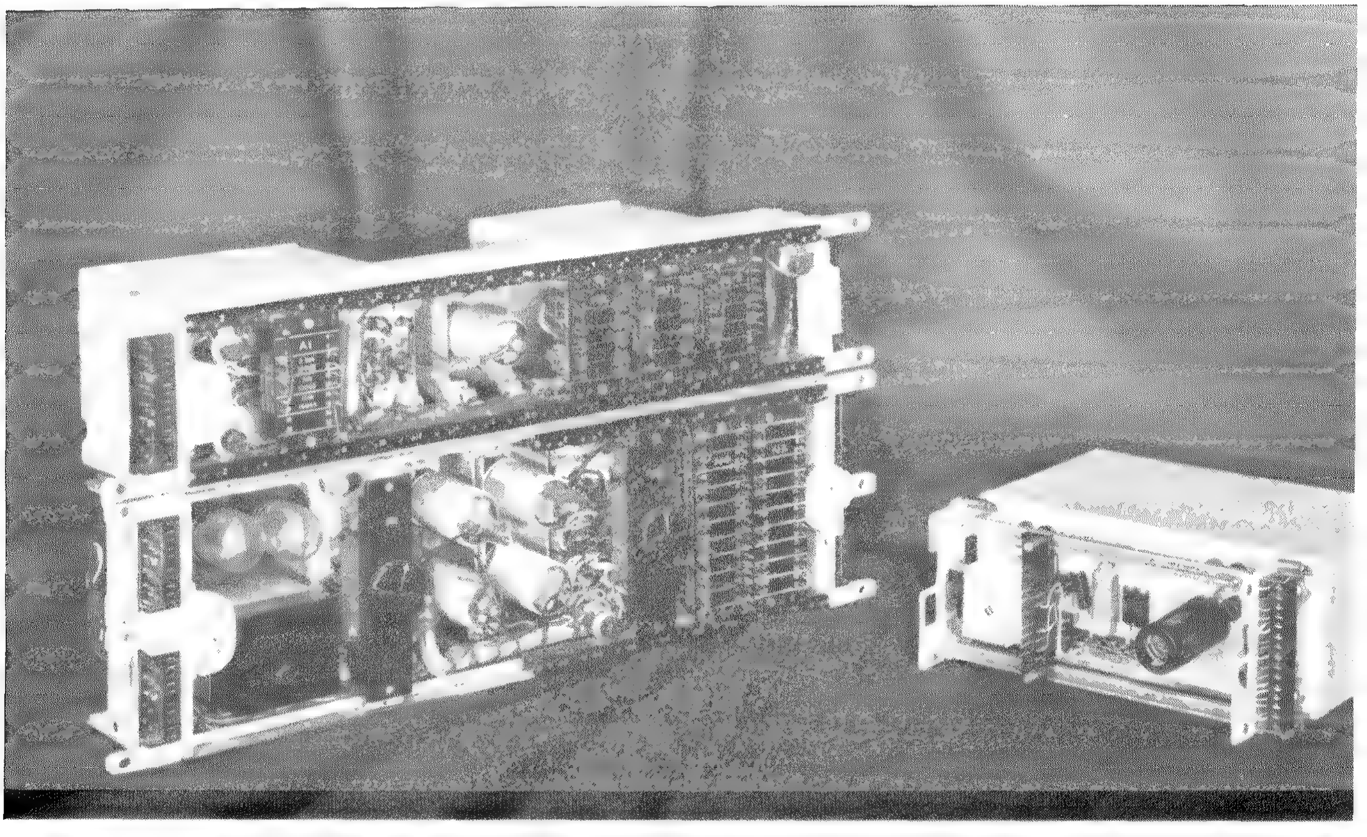


Figure 5—Equipment for *TF1* frequency-shift-modulated voice-frequency telegraph for one duplex channel end. At the left the lower unit is the detector and the upper panel accommodates the oscillator and modulator. At the right is the equivalent equipment in the *TF3* design.

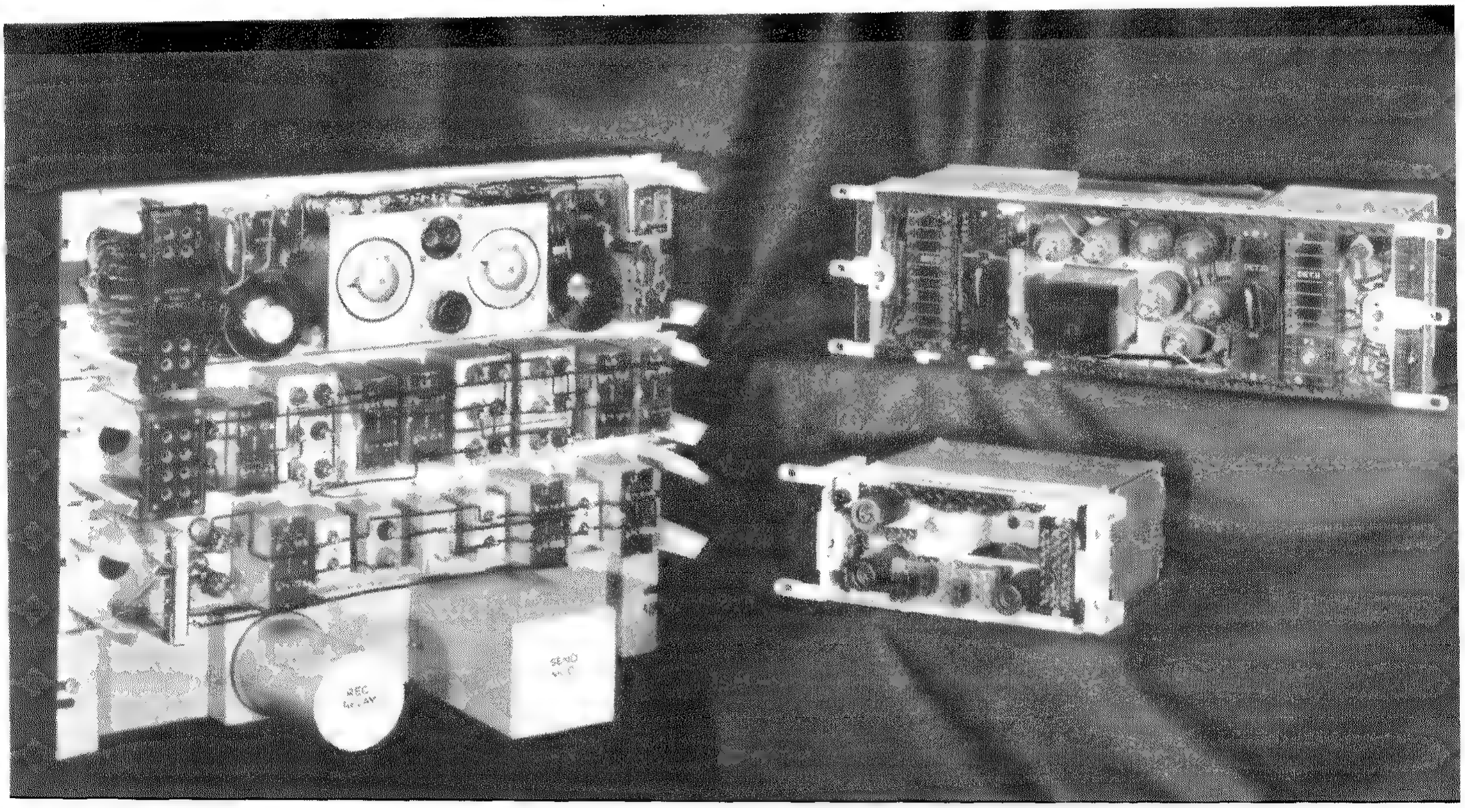


Figure 6—Comparison of three designs of amplitude-modulated equipment for one duplex channel end of a voice-frequency telegraph system.

24 are also derived by a similar group-modulation process; the reason is to ensure the high standard of performance of the equipment over its full temperature range. If produced by direct modulation the stability of the resonant units of channels 19 to 24 would approach their permissible limit. It has therefore been considered advisable not to use direct modulation of these channels for the main applications of this system, but only for special cases where a slightly lower standard of performance can be accepted.

The high standard of performance achieved by the *TF1* system has been maintained in the development of the *TF3* system, and there are considerable advantages with this latest system. Because transistors have been used to replace hard valves the power consumption has been reduced to only 1 watt for a single channel. The use of low-voltage supplies, together with the small powers required in the transistor circuits, enables smaller components generally to be used.

2.4 Mechanical Features

Modern techniques in the mechanical construction of panels, rack sides, and components in conjunction with improved circuits have enabled the cost and size of voice-frequency telegraph systems to be reduced by a most marked extent. A comparison between various amplitude-modulated equipments is shown in Figure 6. The equipment to the left was required for one channel end incorporating a static modulator, send filter, receive filter, detector, and receive relay for a system[7] produced about 1936. Such a system of 18 channels required 5 bays occupying a volume of 10 feet, 6 inches (3·2 metres) by 1 foot, 8·5 inches (0·52 metre) by 1 foot, 3 inches (0·38 metre). On one of these bays was mounted a multifrequency motor-generator, from which were obtained the various channel tones, and which could serve up to 180 channels if required.

The equipment in the upper right corner of Figure 6 is again a complete channel end. In this case an oscillator is provided on the panel to generate the particular channel tone, though the tone could be derived, as previously, from a multifrequency generator. This type of equipment is known as *TA2*, and was designed about 1949. It has become a standard system of the British Post Office, to whom it is known as Type IV equipment. Twenty-four such channels can be mounted on two rack sides of design known

as new equipment practice (NEP).[10,11] These two rack sides when fitted back to back occupy the same space as one double-sided bay of the older design. The small panel to the lower right of Figure 6 is a complete channel end of a *TA4* system; 24 such channels mount on only a single new-equipment-practice rack side occupying a volume of 9 feet (2·74 metre) by 1 foot, 8·5 inches (0·52 metre) by 8·5 inches (0·22 metre).

The general reduction in size of these equipments was brought about by the availability of smaller coils, capacitors, and valves, and improved metallic rectifiers. The mechanical technique at the time of the design of the *TA2* equipment was to leave coils and capacitors individually unprotected from atmospheric conditions, but to put them, together with other components requiring protection, inside common hermetically sealed cans. These cans with their associated valves and other components were mounted on panel frameworks designed to slide into die-cast guides mounted on a light rigid steel rack side framework. The further reduction in size of the *TA4* equipment was by simplified and improved circuit design together with a new mechanical design that employed even smaller coils and capacitors, with resonant units individually sealed in cans of simple but efficient design. The type of valves used helped the general reduction; type *B7G* replaced the octal valves. The simplified circuitry required only three valves, one of these being that of the channel oscillator.

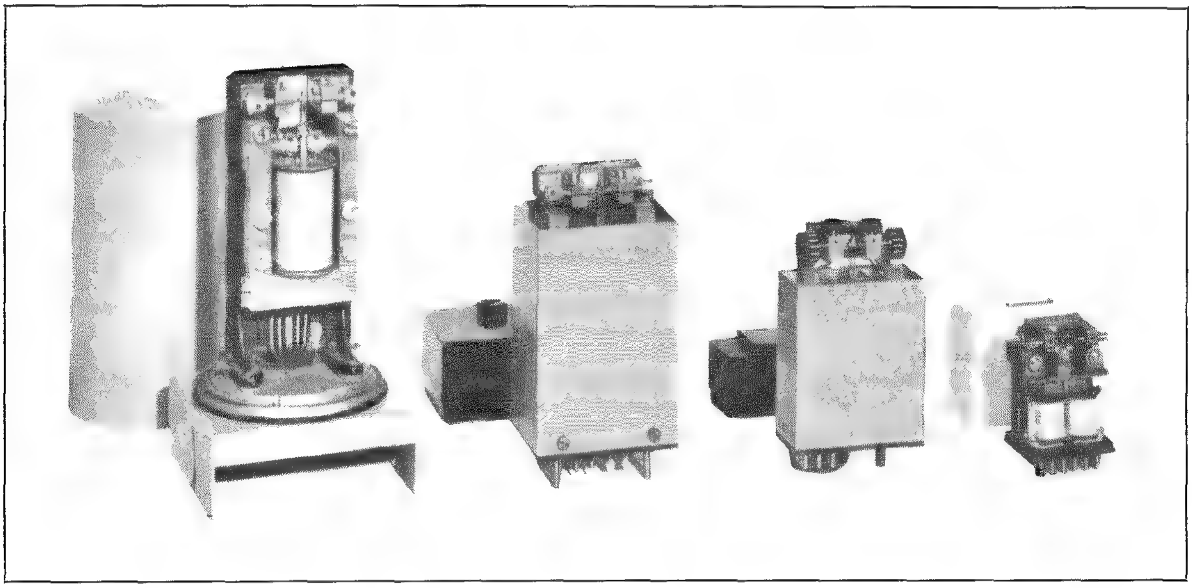


Figure 7—Four progressive designs of telegraph relays.

Frequency-shift-modulated equipments are compared in Figure 5, and just as marked reduction is shown as for the amplitude-modulated equipments. The two panels to the left of the picture constitute one channel end of the *TF1* system. These panels occupy five units of new-equipment-practice rack side, whilst the panel to to the right is for one channel end of the *TF3* system; it occupies only one unit of rack side. This reduction is not so much due to simplified circuit design as to the use of individually sealed resonant units and coils, to improved mechanical construction, and above all to the use of transistors in the place of hard valves.

2.5 Telegraph Relays

One important component in all telegraph systems is the telegraph relay, used to repeat the

[10] F. Fairley, R. J. M. Andrews, and A. C. Delamare, "Improved Equipment Practice Reduces Size of Telephone Transmission Systems," *Electrical Communication,* volume 27, pages 21–38; March, 1950.

[11] E. T. C. Harris and C. J. Spratt, "Improved Form of Mechanical Construction for Transmission Equipment—51-Type Construction," *Post Office Electrical Engineers Journal,* volume 51, pages 197–201; October, 1958.

detected alternating-current signals and to isolate the detector and amplifier circuits from the direct-current receive leg circuits. These relays also have progressed, as is shown in Figure 7. The order from left to right is the *4121* relay used on amplitude-modulated systems up till 1949, the *4148* relay for the *TA2* and *TF1* systems, the *4192* relay on *TA4*, and lastly the *4199* relay for the *TF3* equipment.

It is doubtful whether a further reduction in size of telegraph relays below that of the *4199* will be of much use in practice, since there is always the maintenance problem to be considered. If the physical dimensions of the relay become too small, it will require a highly skilled mechanic to maintain the relay, unless of course a sufficiently robust device is developed that needs negligible maintenance and is sufficiently cheap that at the end of its useful life it, or at least the worn part, could be discarded and easily replaced by a new one.

A modern competitor for the replacement of the electromechanical relay is, of course, the transistor. At the present stage of development and cost, these devices are not really a commercial proposition as a direct replacement for the conventional telegraph relay for the majority of its applications. However, the time can be foreseen when this replacement will occur.

3. *Miscellaneous Systems and Equipment*

3.1 High-Frequency Radio Telegraphy

Multichannel voice-frequency telegraph equipments of the *TF1* type have for some time been used on high-frequency radio circuits. To combat the effects of selective fading special diversity combining panels were provided, such that two telegraph channels, each carrying the same information but received either in space diversity or frequency diversity, could be combined to give one common output. Though these equipments have given satisfactory service, it has been found desirable for the frequency deviation of each telegraph channel to be greater than the existing ±30 cycles per second. Systems have now been developed from the *TF3* equipment with 340- and 170-cycle-per-second spacing of channels, the deviations being ±85 and ±42·5 cycles per second respectively.

The design of the diversity combining equipment is based on the principle of square-law combination.[12] It can be shown theoretically and practically that if two or more channels carrying the same telegraph information are combined in such a way that each contributes to a common output in proportion to the square of its relative received amplitude, then a distinct signal-to-noise advantage can be obtained. The 340-cycle-per-second system using dual diversity is recommended for start–stop telegraph signals at speeds up to 75 bauds and for synchronous transmission up to 100 bauds.

3.2 Submarine Telegraphy

There has been a rapid increase in the number of submarine-cable telephone systems in recent years, and this has resulted in a large increase in the number of international voice-frequency telegraph connections. The voice-frequency telegraphs so used are of the standard land-line type, and a notable case is the transatlantic telephone cable between Great Britain and America. One of the speech circuits to Canada is used continuously for 24 channels of frequency-shift-modulated voice-frequency telegraph equipment of the *TF1* design. These channels were planned initially to operate at 50 bauds and possibly 75 bauds; in fact, some are being used at speeds in excess of 80 bauds.

The vast majority of submarine telegraph circuits still use, and will continue to do so for a very long time, direct-current transmission. The efficiency of existing circuits has been improved by the insertion of a submersible amplifier[13] or, more accurately, part of the receiving equipment, at a point in the sea sufficiently remote from interference picked up in the shallow shore end of the cable. The improvement thus obtainable, although reduced by the fact that a repeatered cable may no longer be operated duplex, is appreciable, and it is customary to introduce such a repeater at each end of the cable with switched by-pass arrangements to permit operation in either direction according to traffic requirements.

[12] H. T. Prior, British Patent 709 793; June 17, 1952.

[13] C. H. Cramer, "Submerged Repeaters for Long Submarine Telegraph Cables," *Western Union Technical Review*, volume 5, pages 81–91; July, 1951.

3.3 High-Speed Data Transmission

As the demand for high-speed data transmission grows it is expected that these basic designs of frequency-shift-modulated systems will be extrapolated to provide any reasonable speed of transmission required over normal telephone circuits, whether the mode of transmission required is a parallel or series mode. Such a system has been developed purely for experimental purposes; this was designed to transmit a message at an aggregate speed of 1750 bauds in a parallel mode over seven 250-baud voice-frequency channels, each of 400-cycle-per-second bandwidth. Its application was of a particular type, and built into the telegraph system were the master speed controls and means to maintain synchronism from the actual message signals being transmitted between the two terminals.

3.4 Start–Stop Regenerative Repeaters

Despite the more widespread use of frequency-shift-modulated telegraph systems with their improved performance, it is still often necessary, to enable long telegraph circuits made up of several links in tandem to function satisfactorily, for the signals to be regenerated and repeated at some point in the circuit. The necessity for regenerators is even more marked on radio-telegraph circuits. Such a regenerator[14] has been developed using transistors and is illustrated in Figure 8. This is an interesting example of the practical application of transistors to digital circuits, and shows how a large amount of this type of circuitry can be mounted within a small space. No coils have been used in the design of this regenerator, the basic bricks being transistor binary circuits together with rectifier gating circuits. The regenerator with all the features found necessary for radio working; that is, adjustable false-start rejection, automatic insertion of a missing stop element, and accurate retransmission of a long space condition, is mounted within a unit 7 inches (18 centimetres) by 3·5 inches (9 centimetres) by 8 inches (20 centimetres). Employing one control oscillator for each 12 regenerators, 36 regenerators and their power supplies can be mounted on a 9-foot (2·74-metre) new-equipment-practice rack side.

[14] W. F. S. Chittleburgh et al, British Patent Application 20 160/56; June 29, 1956.

4. *Conclusions*

The demand for telegraph transmission and the development of such equipment have, in the past, been continuous. Although telegraphic traffic as such has tended to decline in recent times, a new demand has grown up. This is data transmission, which needs the same, or similar equipment. Signalling speeds are being requested that require for transmission only the narrow bandwidth of a standard voice-frequency telegraph channel or that may demand the full facilities of a complete telephone channel. These higher speeds call for new systems but, in most cases, the basic design problem is similar to that for voice-frequency telegraphy. The fundamentals of design that were evolved in the past, can be and are now being applied to these new data-transmission systems.

5. *Acknowledgement*

Thanks are extended to my many colleagues who have helped in the preparation of this article.

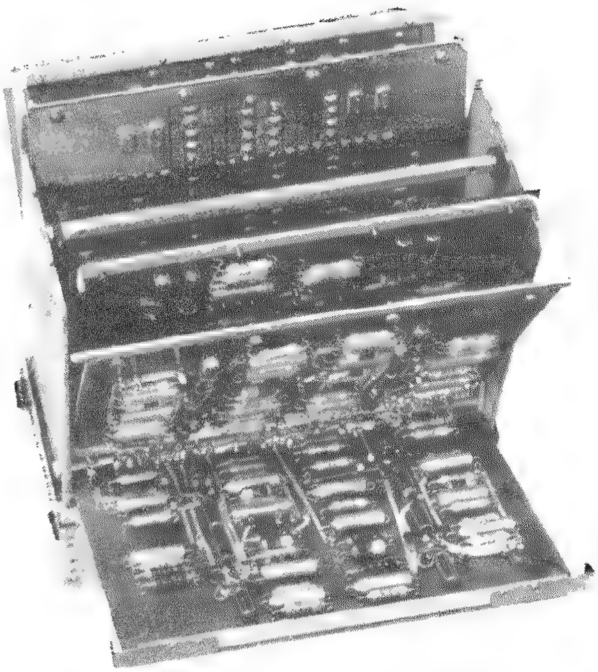


Figure 8—Start–stop-telegraph regenerative repeater using transistors.

Page Printer LO-15-B

By J. AUGUSTIN
Standard Elektrik Lorenz AG; Stuttgart, Germany

ACCEPTANCE of the model-*LO-15* page printer in Germany is attested by the fact that one of every two teleprinters in that country is of this type. Recently, improvements have been made in its design and are incorporated in the *LO-15-B*. The new teleprinter provides key-type controls for the answer-back unit, an all-blank repeats operation, a transmitter–distributor, and a reperforator. In addition, the operating noise has been substantially reduced by minimizing all openings in the housing through which sound generated in the interior of the case might escape.

1. General

The new printer is equipped with a monitor that automatically stops its operation if there is no paper left on the roll. It will resume operation only if paper has been inserted. This prevents loss of messages that could result with unattended operation in answering a call at a time when the machine was out of paper.

The direct-current power needed for operation is provided from the alternating-current mains, which may deliver 110, 127, or 220 volts. The power unit is mounted within the teleprinter housing.

The page printer may be installed in the conventional wooden console or in a new table-type metallic housing. A built-in fluorescent lamp provides uniform glare-free illumination of the printed material. Provision is made to accommodate attachments such as a reperforator or a transmitter–distributor within the hood. Special

Figure 1—*LO*-15-*B* page printer with built-in reperforator and tape transmitter mounted on a desk.

message blanks in folded strip form are provided for by guides to direct the paper to and from the platen and a base designed to collect the folded strip of paper.

2. *Transmitter–Distributor*

The new transmitter–distributor is designated the *LS-424B*. The perforated tape may be inserted from one end or at any other point along the tape. For convenience, the tape travels parallel to the front edge of the unit and the small plate that holds the tape in place may be released by pressing a button. The transmitter–distributor disconnects itself automatically if the tape is not in operating position. It also stops if the connected distant subscriber starts to transmit.

The transmitter–distributor is stopped whenever the mode of operation is changed. If tape is being run through the transmitter to produce a printed copy and a call is received, the stoppage due to the change in mode prevents the taped material from being transmitted to the calling subscriber.

Control keys are provided to operate or stop the transmitter and the perforator; the latter is electrically connected to the page printer by plugs and jacks.

3. *Reperforator*

The reperforator *ELO-514* has also been equipped to perform some additional functions. It can be switched on or off by the distant subscriber through the transmission of certain code combinations. When the tape is about to run out, both optical and acoustical warning signals are given.

A tape monitor has been added that will disconnect the reperforator if it has no tape. If a call is received, the teleprinter is switched to reception without reperforation. If the caller insists on reperforation by transmitting the designated code combination, the transmitter will be switched off entirely to indicate to the caller that there is no tape in the reperforator.

The reperforator is switched on and off by control keys. Special keys are also provided to eject the tape and to retract the tape to erase an error.

In other details and ratings, the *LO-15B* is the same as the previous model.

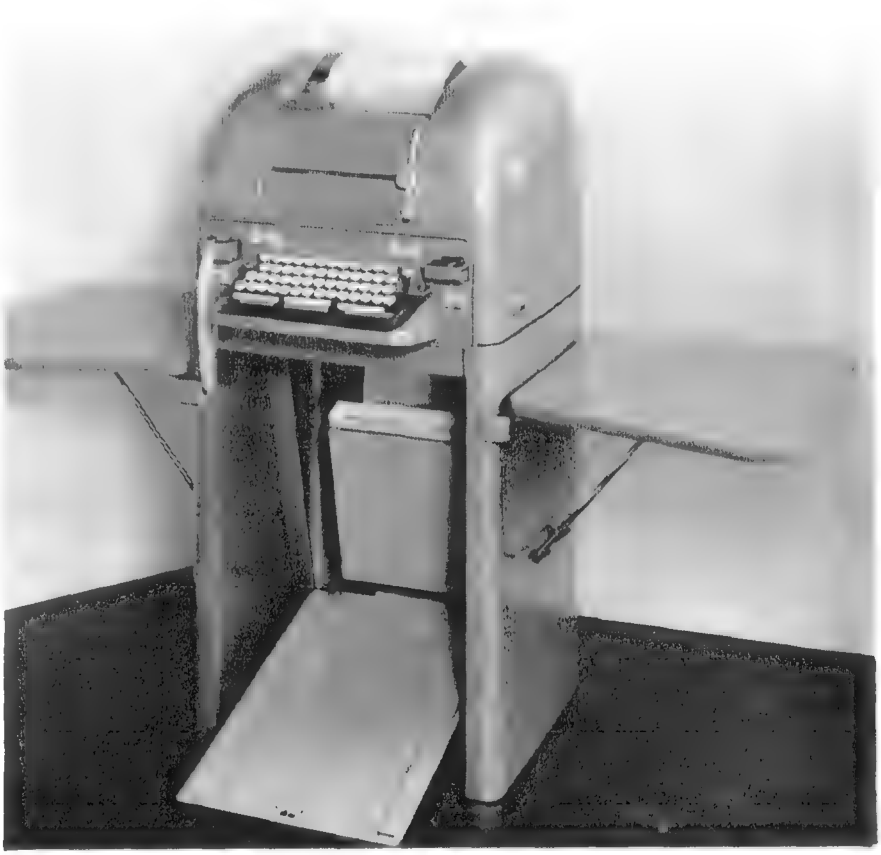


Figure 2—Page printer mounted on a pedestal showing magazine for handling message blanks in folded strip form mounted within the pedestal.

High-Speed Tape Perforator SL614

BY J. AUGUSTIN

Standard Elektrik Lorenz AG; Stuttgart, Germany

WHEN PERFORATED tape is used for permanent storage in data-processing systems, bookkeeping machines, and computers, the perforator used as an output device must have a much-higher speed than the conventional perforators used in teleprinter service.

High-speed perforator *SL614* (Figure 1) has been constructed for such high speeds, of which the following are presently available: 25, 31.25, 40, and 50 characters per second.

Tape can be perforated in any of the 5- to 8-unit codes. All varieties of this machine can be equipped to perforate two identical tapes if required. (A special design for perforation of card edges or McBee cards is under development.)

The high-speed perforator can be adapted to any of the above modes of operation at any time by simply changing the gears and control-magnet systems.

The main units of the machine are: the motor and drive, a set of contacts that generate orders and synchronizing pulses for the associated data-processing system, the tape reels with signaling contacts, the perforating mechanism with the tape driving system, the control magnets,

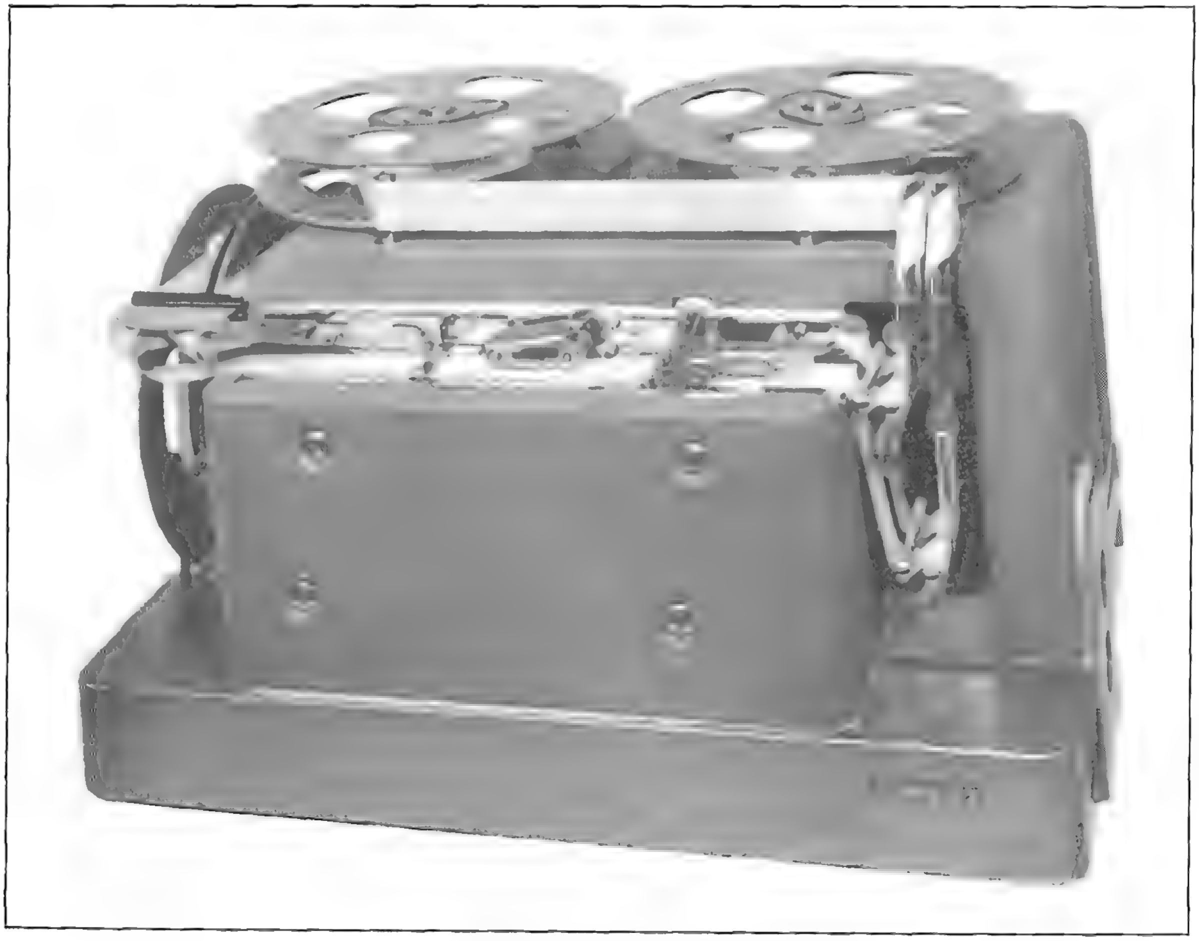


Figure 1—Front view of high-speed tape perforator.

the coupling magnet system, the revertive-signal contacts, the take-up reels, and the control relays. An additional cam-switching unit can also be mounted.

The standard equipment of the *SL614* comprises a universal motor for 220 volts at 50 cycles per second; the speed is regulated by a governor. If required, motors of other types (shunt-wound direct-current or asynchronous or synchronous motors for 50 or 60 cycles per second) can be provided. The power consumption of the drive is 200 volt–amperes. Rated speed is reached within 2 seconds after the motor is started.

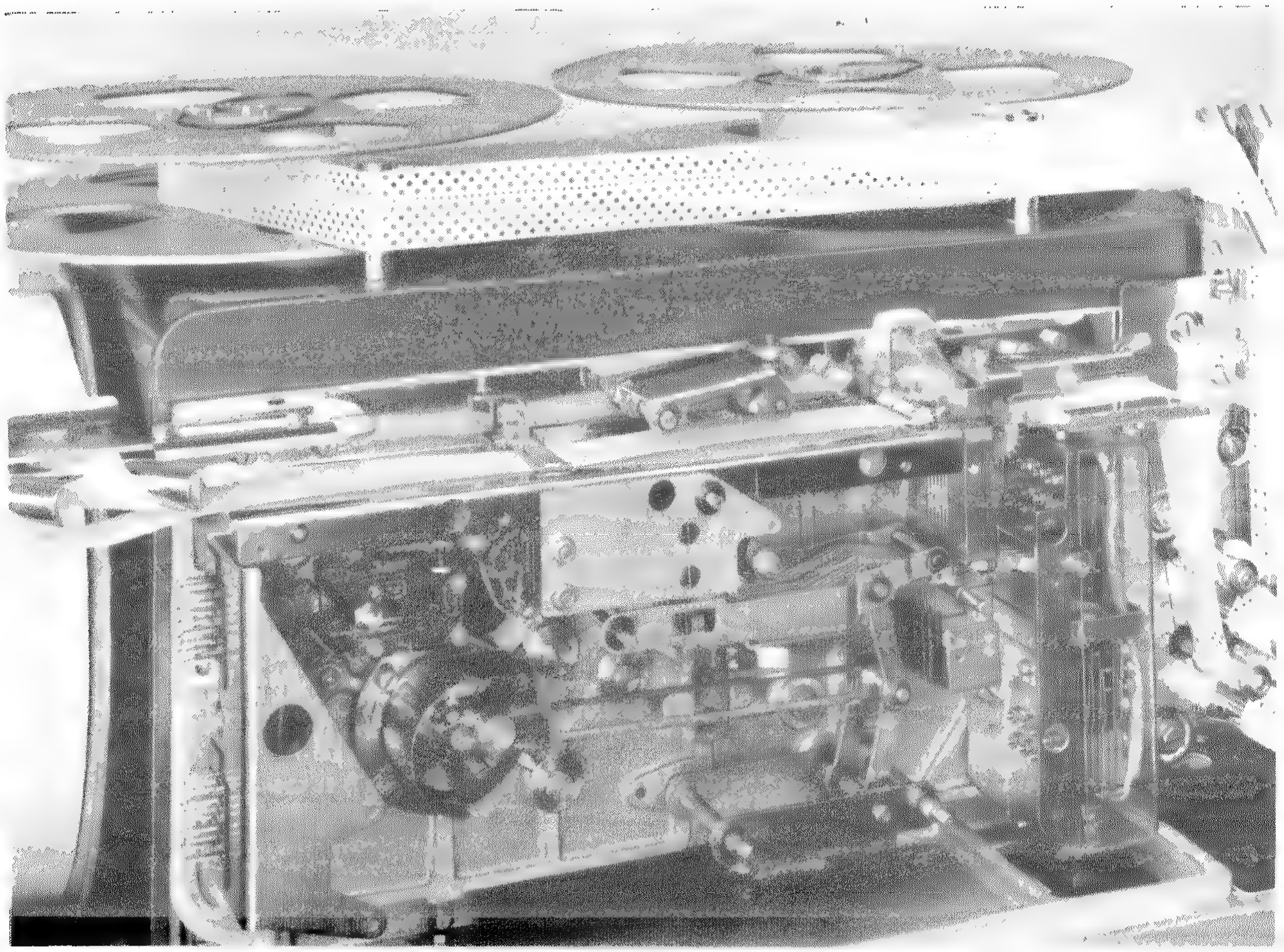
There are 7 control contacts mounted on the main-shaft cam disks. The first contact controls the coupling processes within the machine and the other 6 contacts generate order signals for the associated data-processing system. Cam disks for these contacts are supplied according to the particular application. The contacts are designed for bounce-free operation and an accuracy of ± 1 millisecond. If more than 6 contacts are needed, the cam-switching unit mentioned above is added.

The left side of the perforator is equipped with journals holding the two reels of nonperforated tape. Each reel is monitored by a feeler lever operating a contact to actuate an alarm to warn the operator when the reel is nearly empty.

The tape is led around the corner to the tape guide, the tape-checking lever, perforating head, and to the tape driving system. The tape-checking lever operates a contact signaling whether or not tape is moving properly. This signal can control the output circuit of the data-processing system so that there is no output if tape is not present for perforation. The tape drive pulls the tape from the reel at the proper rate through the perforating unit.

At the output of the perforating unit, tape is pulled character by character by a sprocket wheel to go then to the take-up reels. Mounted on the same base plate is the coupling unit (Figure 2) connecting the perforating unit and the tape drive to the motor during the perforating process. The perforating mechanism comprises a gearbox of bell-crank levers operating the individual perforating pins. This system is controlled by intervening armatures of the control magnets so

Figure 2—High-speed tape perforator with cover removed.



that each perforating pin will operate only when its associated control magnet is energized. The feeding-hole pin operates at every step. At the moment of perforation, the revertive-signal contacts operate; they are interlocked with their associated perforating pins.

After passing through the perforating unit and pin-wheel drive, the perforated tapes are taken up on the reels on top of the machine. These reels are driven by the motor through separate friction clutches, ensuring tight winding of the tapes. Pins between the tape drive and the takeup reels guide the tape and act as brakes inasmuch as the pin wheel is not loaded by the pull of the reels.

Relays facilitate control of the perforator by the associated data-processing system. One relay switches the perforator on; the other operates when the tape runs out. Two relays are interlocked so that tape replacement is possible only when the machine is switched off and accidental depression of the control keys does not interrupt perforation.

The individual sub-units of the perforator are interconnected by plugs and jacks for convenient replacement and servicing. An interlock circuit ensures that the machine will not operate if any of the plugs are disconnected. The paper chaff box mounted at the right-hand side can contain the chaff from two perforated tape rolls.

The high speed of the *SL614* required special design for some units. Thus, the armatures of the control magnets are retained mechanically and either released or held magnetically, depending on the state of current. Moreover, the time constants are as small as possible; this necessitates higher power consumption in the control circuits.

Materials of best choice enable operation for long periods with minimum maintenance. Therefore, the life of the perforator depends greatly on the frequency of on-and-off switching of those parts involved in perforating. Hence, discontinuous modes of operation, where the output is individual characters, should be avoided. Operation should preferably be stopped only when the entire message is finished. Use of unsuitable paper can result in premature wear of perforating pins and dies.

The motor is conventional; its leads are equipped with fuses and are free of electrical interference. The current and voltage depend on local conditions.

All control terminals are brought to contacts on the rear side of the machine; from where connections can be made to the associated data-processing system. The high-speed perforator has no telegraph-current sources of its own.

The machine is 20.5 inches (520 millimeters) wide, 12.6 inches (320 millimeters) high, and 14.4 inches (365 millimeters) deep; it weighs 44 pounds (20 kilograms). Available speeds are 25, 31.25, 40, and 50 characters per second in codes from 5 to 8 units. The power requirement is 200 volt–amperes and suitable equipment can be provided to operate at the voltage and frequency of the power mains in the locality of use. The direct-current control power that must be supplied by the associated data-processing system is 60 volts at 1.2 amperes maximum.

Teleprinter for Reliable Transmission of Numbers*

By J. AUGUSTIN

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PRIVATE and public administrations often transmit numerical teleprinter information. When straight text is transmitted, an error consisting of a wrong letter is most often quickly apparent and the correct letter can be deduced from the context. It is quite different with numbers. The receiving subscriber usually has no means of perceiving erroneous numbers or replacing them by the correct numbers. Errors can be caught by complete retransmission of each individual message; visual inspection of the two versions of the message will reveal any errors. However, this method is rather costly, especially where much information is handled.

There was good reason, then, to develop a teleprinter for reliable transmission of numbers. Compatibility of such a machine with the public telex systems is of particular importance; in the majority of cases, maintenance of a private network is too expensive.

The machine developed has three shift positions instead of the conventional two. The first two shifts comprise the usual lower- and upper-case characters of any ordinary teleprinter. The third shift provides reliable transmission of 15 characters: The numbers 0 through 9, plus sign, minus sign, space, carriage return, and line feed.

Of the 32 possible combinations in the 5-unit code, only those 15 were selected that are characterized by 1 and by 3 marks in the 5 positions. Hence, at least 2 elements of a combination have to be wrong to cause an error. In principle, combinations having 2 and 4 marks in the 5 positions could also have been selected. However, the combinations for carriage return, space, and line feed also must be used in the first and second shifts. In telegraph systems operated according to Comité Consultatif International Télégraphique et Téléphonique recommendations, these three combinations are of the group having 1 mark in the 5 positions; this decided the case in favor of the 1-mark and 3-mark combinations.

Figure 1 shows a teleprinter providing reliable

Figure 1—Teleprinter for reliable transmission of numbers. The set of black keys used in the third-shift position is visible above the conventional keyboard.

* Originally published under the title "Fernschreiber zur gesicherten Ubertragung von Ziffern" in *SEG-Nachrichten*, volume 6, number 1, pages 58–59; 1958.

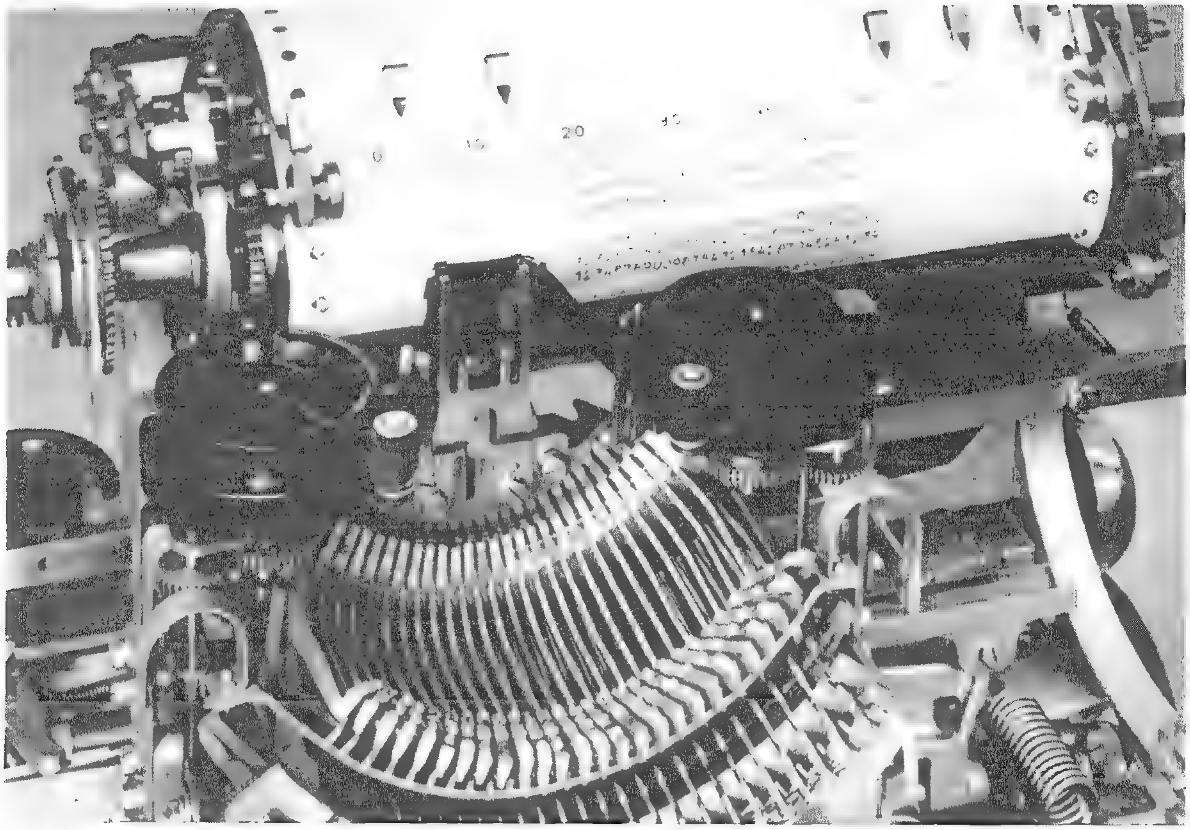

Figure 2—View of the type bars with three characters each and the teleprinted message in which the italics denote the high-reliability numbers.

transmission of numbers. The conventional keyboard has white keys. Above it are another 12 back keys for 0 to 9, plus, and minus. A locking bar releases the white keys only in the first- and second-shift positions; the black keys only in the third-shift position. Accidental operation of the wrong keyboard is thus precluded.

Each of the type bars carries three characters. (Figure 2). The position of the platen in any of the three shift positions determines which of the three characters of a type bar will print.

The numbers transmitted in the third shift are clearly distinguished from the conventional numbers by printing in italics. All type bars not equipped with the third-shift characters have an asterisk in the third-shift position. This asterisk signals that an erroneous number-combination has been received; as a result, no special electrical check measures are necessary.

If its only current impulse is dropped, each of the 1-mark-in-5-position combinations becomes combination *32* of International Telegraph Alphabet *2*; that is, the combination releasing the third shift. Since the receiver could not determine whether this code is transmitted on purpose or is just a faulty combination, a locking device has been attached to release the third-shift key only when the carriage position is at the beginning of a new line. In addition, a special type bar has been provided to print an asterisk each time the combination for the third shift is received (except at the start-of-line position). Thus, any faulty character converted into combination *32* can be identified.

The above arrangements offer the same protection against mistakes as message repetition in a conventional teleprinter circuit; however, they save the considerable cost that would be incurred by the doubled transmission time.

As mentioned before, this teleprinter operates in the conventional manner when in the first or the second shift and can be employed to communicate with every subscriber in a public telex network.

Teleprinter Synchronizing Set SYZ-634

By W. SCHIEBELER

Standard Elektrik Lorenz AG; Stuttgart, Germany

START–STOP teleprinters of any description may utilize the *SYZ-634* attachment. It is designed to bridge a time of disturbed transmission by supplying locally generated start and stop signals to the receiving teleprinter, thus maintaining the rhythm of the teleprinter transmission. It is therefore particularly useful for radio transmission and for the transmission of enciphered messages. Its application covers the whole field of automatic transmission in which the transmission speed is normally uniform.

There are two reasons for restoring start–stop pulses lost by disturbed transmission paths. One of these concerns the case of plain-language communication in which loss or distortion of one or more teleprinter signals and the associated start–stop pulses causes not only incorrect reproduction of that particular character but may also distort subsequent characters that were correctly transmitted.

If the receiving page printer fails to receive one or more stop pulses, the selector mechanism will run through and will not be synchronized until several characters later. This action is prevented by the use of the synchronizing set *SYZ-634*, by means of which the first character correctly transmitted after an interruption is also reproduced correctly. As a result, the number of faulty characters printed in conventional operation is considerably reduced by the *SYZ-634*.

The other case is that of enciphered messages in which the consequences of start–stop pulses being lost in transmission and the receiving deciphering set losing synchronization with the transmitting encoding set are even more serious than in plain-language communication. This is because the enciphering at the transmitter and deciphering at the receiver are on a character-by-character basis and a character lost in transmission throws all following characters out of proper deciphering order. The rest of the message, therefore, becomes illegible even though the following transmitted characters are correctly received.

The synchronizing set is designed to bridge interruptions of at least 3 seconds caused by distortion; this corresponds to 20 characters. It is adequate for the most-unfavorable case of such a disturbance. Bridging times of 20 seconds and more can be achieved for complete interruption of the transmission path.

The most important task of the synchronizing set is to generate start and stop pulses continuously at the correct frequency and phase established by the transmitter. In the synchronizing set, the received start and stop pulses are diverted to a pulse generator and the newly developed pulses are inserted into the received signal in the correct phase to actuate the teleprinter. As long as undistorted characters are received, the pulse generator is frequency controlled continuously by the mark-to-space transitions of the stop pulses. If these transitions are not received due to a disturbance in transmission, pulses continue to be generated by the local pulse generator at the last-received frequency, thus ensuring uninterrupted operation of the associated teleprinter receiver.

As will be seen in Figure 1, the teleprinter characters arriving from the line are applied to a line directional switch that removes the combined stop and start pulses and sends only the character pulses through the teleprinter directional switch to the teleprinter.

The removed stop pulses are differentiated, the negative pulses produced by the leading edges are suppressed and the positive pulses from the trailing edges go to a trigger circuit to control the frequency and phase of new pulses produced by a pulse generator under its control. From the generator, the pulses pass through the teleprinter directional switch to be combined with the character pulses received directly from the line directional switch, the combined signals actuating the teleprinter.

If the teleprinter signals being received from the line are interrupted, no positive pulses will be received by the trigger from the differentiator and line directional switch. Nevertheless, the pulse generator will continue to supply stop and

start pulses through the teleprinter directional switch to the teleprinter. The latter will receive code groups consisting of all marks or all spaces depending on the condition of the line and which corresponds to "letter shift" or "ff". However,

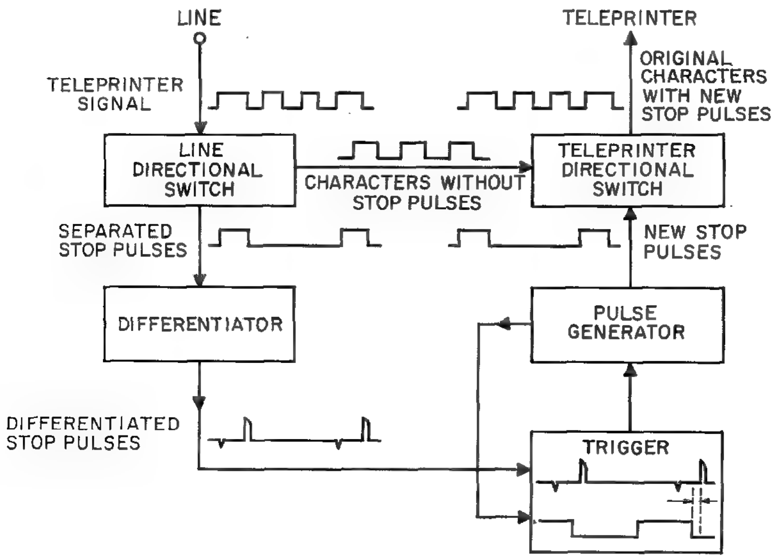

Figure 1—Operation of synchronizing set.

in the case of enciphered messages, for instance, the tapes continue to move synchronously at transmitter and receiver so that, when the disturbance disappears, correct text is again printed.

The principal unit of the synchronizing set is a free-running multivibrator, the circuit of which is shown in Figure 2.

Contrary to conventional multivibrators employing a double triode, two double triodes are used. One pair of triodes *V1* has the function of a conventional multivibrator. Cutoff pulses are alternately applied to the two grids so that one triode is always nonconducting while the other conducts heavily. After a certain time when the grid-circuit capacitors have been charged or discharged, respectively, the state is reversed so that the first triode conducts and the other does not.

The second pair of triodes *V2* controls the frequency of the multivibrator. A slightly positive or negative voltage is applied to the grids of *V2* which are connected together. This voltage also appears across capacitor *C1*. If this voltage is positive, the pulses are lengthened; if the voltage is negative, the pulses are shortened. At zero grid voltage, the pulse duration is 150 milliseconds, that is, it coincides with the duration of a teleprinter character.

The plate currents through *V2* are subject to pulse-shaped alternations similar to those passing through *V1*. However, the cutoff of triodes *V2* is not as complete and uniform as in *V1* but produces a sloping pulse crest. The plate-current pulse variations in *V2* are not controlled by the grids of these triodes, which are

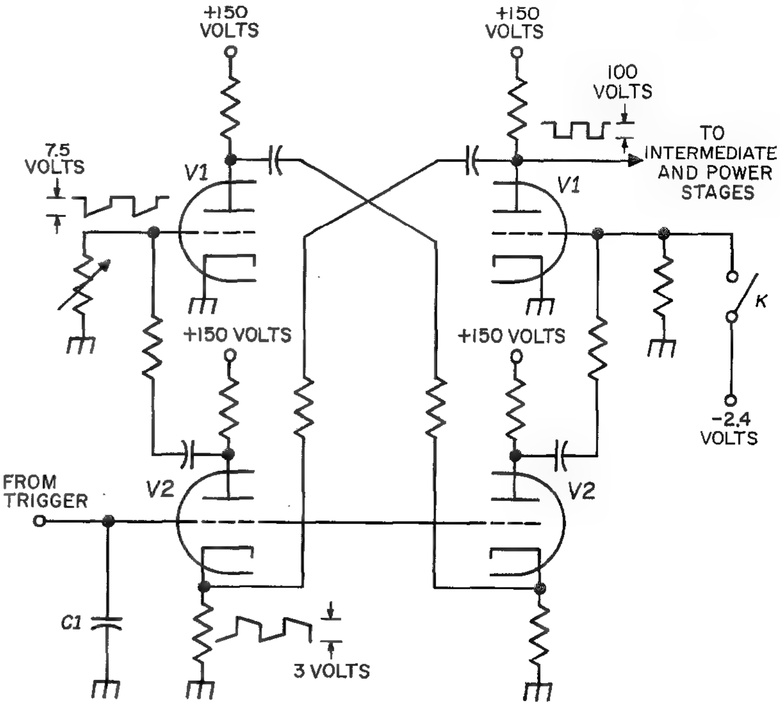

Figure 2—Multivibrator pulse generator. Signals from the trigger circuit to the double triode *V2* control the length of the pulses generated by the two triodes of *V1*. *C1* has a capacitance of 8 microfarads.

connected to the trigger circuit but by pulses obtained from the plates of *V1* and applied to the cathodes of *V2*.

The symmetric square-wave pulses from *V1* are applied to the grid of an amplifier employing a pentode connected as a triode to provide sufficient power to operate the receiving magnet of the teleprinter.

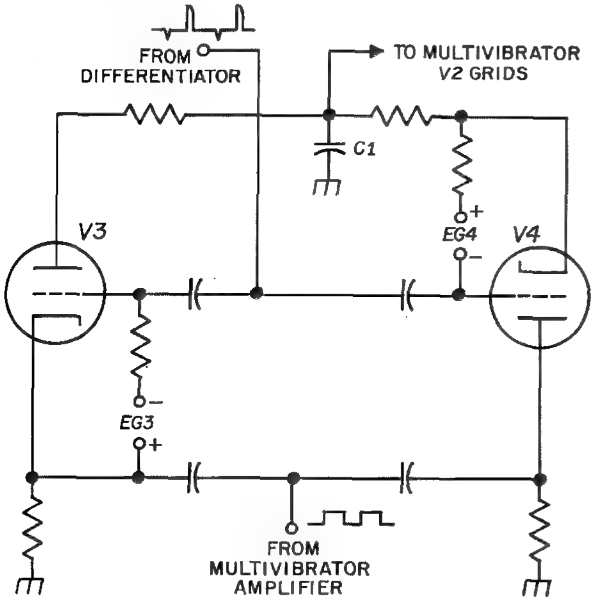

Figure 3—Trigger unit. The bias voltages *EG3* and *EG4* are adjustable.

In addition, the plate-circuit pulses of *V1* are applied through a buffer stage to the trigger unit. The buffer stage is provided to protect the multivibrator from overload.

The trigger unit employs the circuit shown in Figure 3. Its function is to compare continuously the positive control pulses obtained by differentiation from the trailing edges of the stop–start pulses of the received teleprinter code with the timing of the new stop–start pulses produced by the generator. If there is a difference in time between the trigger pulse and the new pulse, the grid voltage of the multivibrator tubes *V2* must be increased or decreased, depending on the sign of the difference. This ensures synchronization between the distant tape transmitter and the multivibrator.

The mode of operation of the trigger unit is as follows. Two triodes, *V3* and *V4*, are connected to capacitor *C1* of the multivibrator tube *V2* in Figure 2. The square-wave pulses from the multivibrator are amplified in an intermediate stage and applied as plate voltage to *V3* and *V4*. When *V3* and *V4* have no negative grid bias, that is, are not operating at plate-current cutoff, the positive half-wave of the pulse can flow only through *V4* and the negative only through *V3*. In this way, *C1* alternately receives positive and negative charges at the pulse-repetition frequency. The multivibrator connected to *C1* then operates somewhat slower during the positive half wave and somewhat quicker during the negative half wave of the pulse.

Generally, however, *V3* and *V4* are blocked against any flow of current by high grid bias voltages. Capacitor *C1* receives no charge and retains its voltage. Only when the positive pulses from the differentiator are applied to the grids of *V3* and *V4* do both tubes conduct for a short time, during which the charge across *C1* can vary. The sign of the charge variation depends on the phase of the multivibrator pulse instantaneously applied as plate voltage to *V3* and *V4*. If it is positive, a current pulse will flow through *V4* and charge *C1* positively. If the phase is negative, the current will surge through *V3* and produce a negative charge on this capacitor. The multivibrator reacts to this charge variation by a corresponding frequency variation.

The frequency of the multivibrator pulses and of the corresponding pulses of the teleprinter signals differ very little from each other. Hence, several sequential pulses from the multivibrator usually occur so as to overlap completely the short controlling pulses from the integrator. This causes the multivibrator to reduce the length of the pulses and increase the length of the spaces, corresponding to a higher frequency. This process is indicated in Figure 4. Through this action, the multivibrator pulses are gradually changed in their duration until their edges coincide with the narrow positive pulses, whose spacing is assumed to be constant. As soon as the positive pulses fall into the reverse phase of the multivibrator

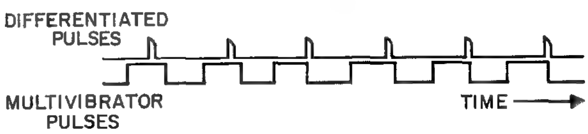

Figure 4—Synchronization of multivibrator pulses.

pulses, the controlling action starts in the opposite direction. After a moderate overshoot, the pulse edge is again pulled towards the positive spike and clamped. The multivibrator pulses show so small a variation about the correct timing set by the short pulses from the integrator that this is not operationally significant.

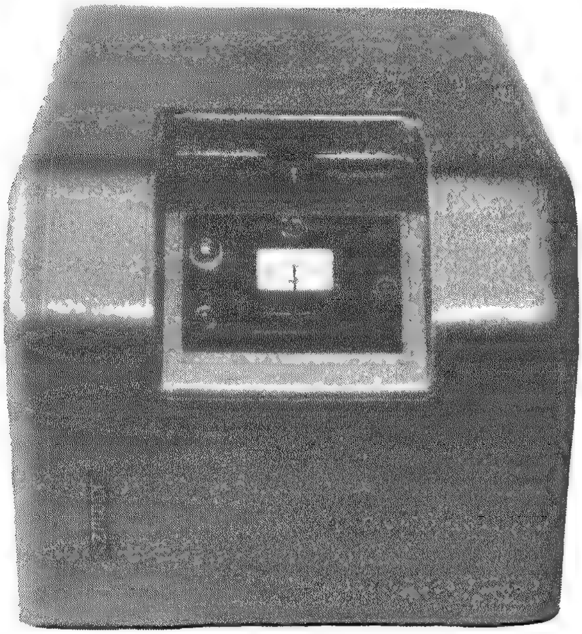

Figure 5—Synchronizing set in housing.

If the teleprinter characters and, hence, the positive spikes do not arrive at the trigger unit, the plate current of *V3* and *V4* is cut off by the negative grid bias; the charge on capacitor *C1* is no longer varied, and the multivibrator oscillates at the last-adjusted frequency. When the teleprinter signals are received again, synchronization will be resumed provided the time difference between the positive integrator peaks and the multivibrator-pulse trailing edges has not exceeded $+20$ or -30 milliseconds. If it has, the control pulses will no longer be applied to the triggering circuit because the line directional switch will have already disconnected the line from the differentiating stage. In this case, no proper deciphering can be expected.

The two directional switches, one connected to the line and the other to the teleprinter, consist of two transfer contacts of a relay operated by a univibrator in the synchronizing set. Operation is such that, as soon as the stop pulse begins, the line is electrically connected to the differentiating unit and, at the same time, the generator is connected to the receiver magnet of the teleprinter. Towards the end of the start pulse, these connections are opened again and the line is connected to directly the teleprinter receiver magnet.

Contact *K* of the multivibrator, shown in Figure 2, is initially closed to place a negative bias on the grid of the output triode of *V1* and stop the multivibrator from oscillating. Hence, the multivibrator amplifier stage supplies continuous current to the receiver magnet of the teleprinter. As soon as the first start pulse arrives from the toll line, contact *K* is opened, the multivibrator begins its first oscillation, disconnects the continuous current to the receiver magnet by cutting off the final-stage tube, and allows the teleprinter to start operating.

The sychronizing set is shown in Figures 5 and 6. It has a meter on its front panel to indicate the voltage across *C1*. If the voltage differs substantially from zero, this is a sign that the fundamental frequency of the multivibrator deviates from the teleprinter frequency. The multivibrator frequency may be readjusted by a control knob on the front panel. The set can be switched to telegraph speeds of 50 and 45.5 bauds.

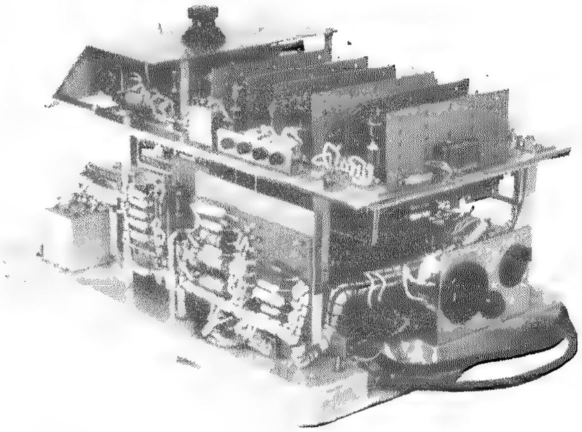

Figure 6—Synchronizing set with cover removed.

High-Speed Gas Tubes for Switching*

By A. H. BECK and T. M. JACKSON

Standard Telecommunication Laboratories Limited; London, England

DURING the past ten years a series of cold-cathode gas tubes has been developed specially for applications in telephone and telegraph switching and for use in computers. A very-brief account of the earlier tubes is given. Two recent advances described in more detail are: A tube designed to pass voice-frequency currents, which is easily incorporated in a switching matrix, and a high-speed triode trigger tube capable of operation at speeds up to at least 1 megacycle per second. The physical obstacles to further progress are discussed.

• • •

For the past decade an extensive programme of investigating the properties and uses of cold-cathode discharge valves in telephone and telegraph switching has been in progress. A short account of the major results obtained is given here.

All the tubes to be described have a common feature:—they use pure metallic cathodes of nickel or molybdenum. This permits the addition of a deionisation agent, usually hydrogen, which greatly increases the operating frequency range. For precise discharge-initiation timing, it is necessary to include a permanent electron source known as the primer. The high work function of the cathode materials precludes use of priming by external illumination and requires an internal priming source. Early experimental work established the principal mechanisms involved in the main discharge; (ionisation coupling, dynamic breakdown, and deionisation) upon which later tube developments were based. This work extended the study[1] of the Townsend discharge by use of modern pulse techniques. Based on these studies, the first counter tube made (1947) was of the progressive-glow-discharge type, based on the variation of ionisation coupling with spacing. This tube had two disadvantages, the multiple glow condition and the need for resetting the tube at the end of its cycle.

The next step was to eliminate the multiple-glow condition; other possibilities were examined. These are shown in Figure 1 and are all based on extinguishing a conducting gap by initiating

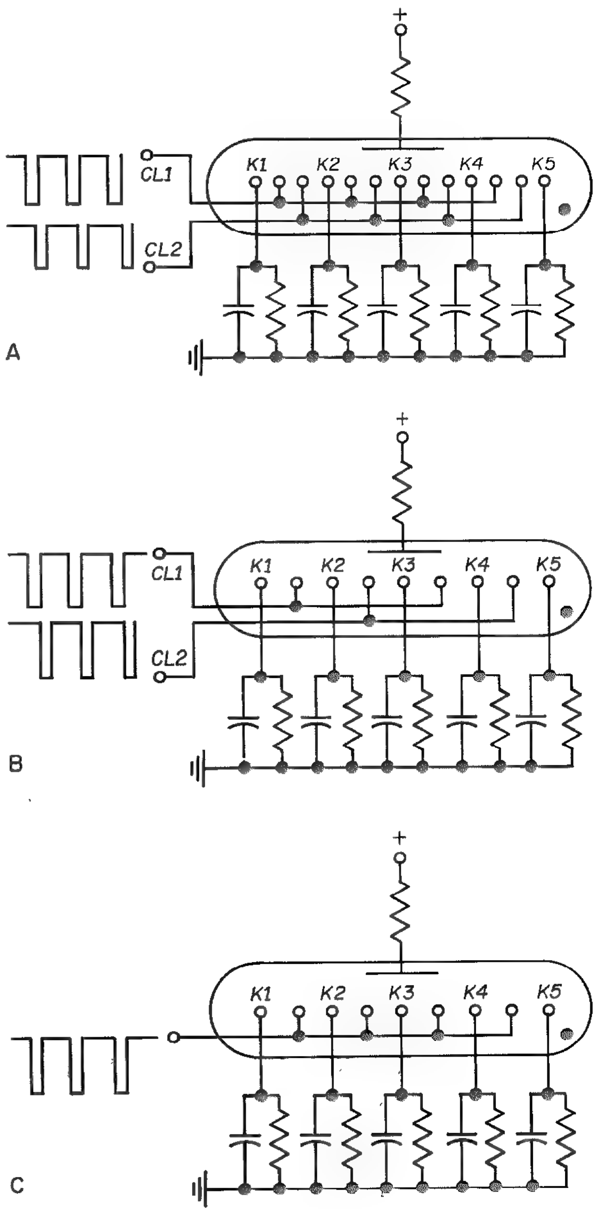

Figure 1—Some electrode arrangements for multicathode gas counting tubes.

* Presented before the German Society of Telecommunication Engineers at Aachen, Germany; February, 1957.

[1] G. H. Hough, Thesis, University of London; 1950.

a new discharge on an intermediate electrode, causing a reduction of the anode voltage below the level necessary to maintain the previous discharge. In type *A* (Figure 1), the pulse on *CL1* extinguishes *K1* and the overlapping pulse on *CL2* transfers ionisation to the proximity of *K2*. Type *B* has input pulses switched alternately between *CL1* and *CL2*, the order of the pulses deciding the direction of glow movement along the array. The arrangement of type *C* was adopted and became the basis[2] of a decade counter tube, the Nomotron.

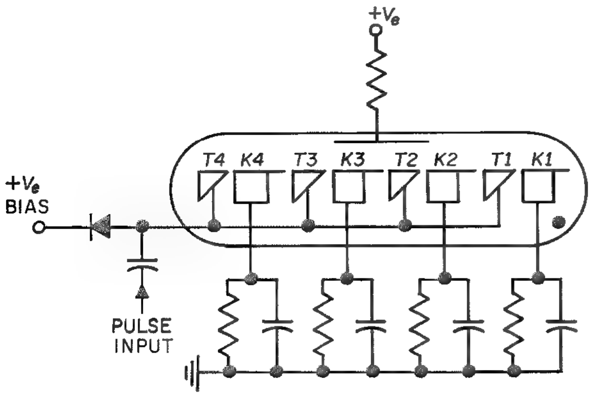

Figure 2—Diagramatic representation of electrode structure and basic circuit connections of Nomotron. The count progresses from right to left.

1. Nomotron Operation

Directional glow stepping in this tube is based on the glow characteristic of a geometrically asymmetric cathode. It is seen from Figure 2 that the cathode has two distinct regions: —the plate and the thin edge referred to as the tail. Apart from the properties of the cathode material and the gas, the voltage required for maintaining discharge also depends on diffusion of charges. This factor is largely controlled by the ratio of peripheral to total area, consequently the maintaining voltage V_m of the tail portion is higher than that of the cathode body and the glow discharge will be more easily maintained on the plate.

Having established this condition, it follows that the transfer electrode adjacent to the glow will be preferentially primed. All transfer electrodes are commoned as shown in Figure 2; the transfer system is directional when the time constant of the cathode circuits is long compared with the input pulse width, because bias remains on the extinguished cathode when another cathode breaks down. The cathode current is nominally 3·5 milliamperes and the tube operates at frequencies up to 20 kilocycles with a negative 10-to-15-microsecond input pulse. Trouble was experienced in some applications of this tube due to changes in characteristics after long periods

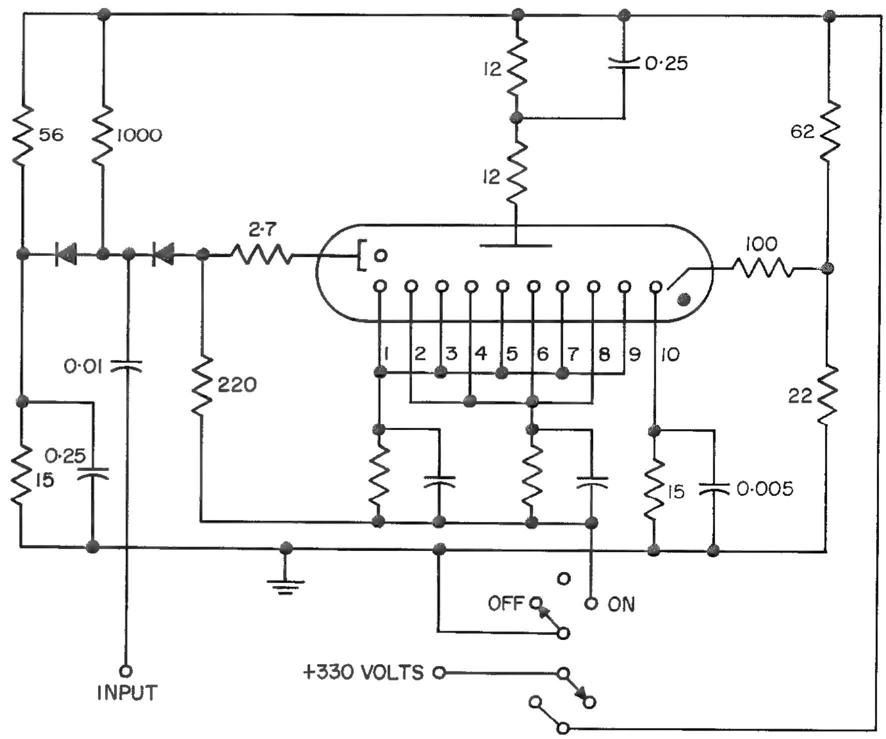

Figure 3—Nomotron circuit for operation up to 5-kilocycle input recurrence frequency. Resistances are in kilohms and capacitances in microfarads. The input is a series of negative square pulses of 120±15 volts amplitude, 16±4 microseconds wide.

[2] G. H. Hough and D. S. Ridler, "Cold-Cathode Glow Discharge Tubes," *Electronic Engineering*, volume 24, pages 272–276; June, 1952.

of conduction to a particular cathode. These changes were attributed to cataphoresis, by which gaseous impurities were concentrated on electrodes adjacent to a conducting electrode. This phenomenon has now been eliminated[3] and satisfactory operation obtained. A complete circuit for operation at input recurrence frequencies up to 5 kilocycles is shown in Figure 3. All 10 cathodes are brought out to enable the tube to be used as a distributor when required.

2. *Trigger Tube G1/371K*

To complete gas-tube counter systems, it was necessary to have trigger tubes covering a similar frequency range and the *G1/371K* trigger tube was developed for this purpose. The salient points follow. Photo-electrons are produced in the trigger–cathode gap by a beam of photons from a priming discharge. To obtain close limits on trigger–cathode breakdown voltage, the dimensions are arranged so that the gap is equal to the physical length of the cathode fall region; that is, at the pressure times distance value for the minimum of the Paschen curve. With the anode–cathode gap chosen, the glow will transfer

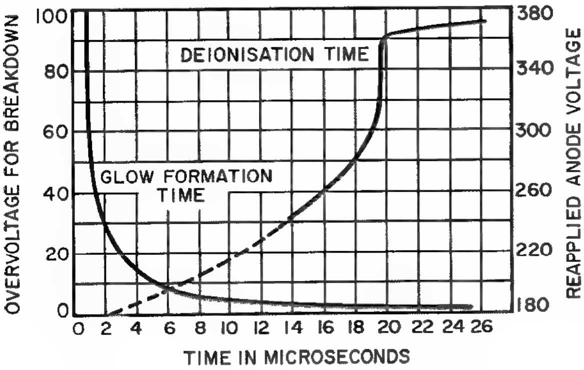

Figure 4—Breakdown voltage and deionisation time for *G1/371K* trigger tube.

from trigger to anode over a wide range of applied voltages. Metallic channels joining the anode and cathode support micas restrict the total diffusion volume and assist deionisation. Figure 4 shows typical breakdown and deionisation time characteristics. The tubes are made to close mechanical limits and are individually tested on most parameters. Large numbers of tubes are in use and have given good performance with little change in characteristics after 25 000 hours of use.

3. *Speech-Gap Tube*

Voice-frequency electronic switching technique requires a device allowing the design of a completely electronic telephone exchange by provision of switched speech paths. The broad requirement for such a device can be stated as a relatively noise-free audio-frequency path of sufficiently low impedance and suitable power-handling capacity to allow the necessary number of series elements in a two-wire circuit to be used without amplification. Also, the static and dynamic characteristics should allow the switching operation to be performed by pulses. Clearly, gas-discharge tubes were capable of performing the switching functions but known tubes were unsuitable for speech transmission.[4] The first task was therefore to examine the transmission characteristics of gas discharges.

Examination of diode gaps of various forms showed the impedance to be made up of resistive and inductive components. The parallel inductance depends on current and frequency in the audio-frequency range. The result was that the impedance was not simply related to the slope of the static characteristic and was of the order of 1000 ohms at practical current levels.

An investigation of the impedance of the various regions of the glow discharge was carried out by means of moveable probes, on the basis that plasma regions of high density should show low resistance. With balanced probes biased to the space potential, hence not conducting current, it was found that the impedance between the probes was high in all regions of the discharge. It was also observed that noise was present in the positive column but completely absent in the Faraday dark space. Further tests made with probes adjusted to conduct current showed that the impedance was high in the positive column when conducting either ion or electron current but in the Faraday dark space, a low impedance was obtained when the probes drew electron current.

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[3] T. M. Jackson, "Improved Circuit for Reliable Operation of Nomotron Counter Tubes," *Electronic Engineering*, volume 29, pages 324–326; July, 1957.

[4] A. H. Beck, T. M. Jackson, and J. Lytollis, "Novel Gas-Gap Speech Switching Valve," *Electronic Engineering*, volume 27, pages 7–12; January, 1955: also, *Electrical Communication*, volume 32, pages 179–189; September, 1955.

3.1 Double-Anode Tube

When the circuit was modified so that all the cathode current was taken by the probes, the main anode of the experimental tube could be disconnected. The path between the two probes (now anodes) was found in this way to be substantially independent of their spacing, provided the anodes were situated on the cathode side of the outer edge of the Faraday dark space. A plot of resistance against pressure is shown in Figure 5 for pure helium at a spacing of 2 millimeters (0·08 inch). The resistance shows a rapid increase when a positive column is formed and a space-charge sheath surrounds the anodes. The static characteristic is illustrated in Figure 6. The central intersection, at which the anode currents are equal, corresponds to no signal input. The slope of the characteristic gives the anode-to-anode (speech-path) resistance. It follows that a cathode current of 20 milliamperes can be switched entirely to either anode by a change of ±1 volt; the modulation is 10 milliamperes. The mechanism of the low gap impedance was established by probe and other measurements that showed the plasma potential following closely the highest anode potential. Consequently, when the potential of one anode is raised, the other anode becomes negative with respect to the plasma and its current decreases. Similarly, when one anode voltage is lowered, the plasma remains at the potential of the higher anode. Figure 7 shows the variation of alternating-current resistance with current, which controls the plasma density and therefore the conductivity. Probe measurements indicate an electron density of 10^9 electrons per cubic centimetre. Figure 8 shows the alternating-current resistance to be independent of frequency. The curves go up only to 20 kilocycles but, in fact, the independence is maintained up to very-high frequencies, as it is solely an electronic effect. The points involved are best considered with direct reference to the design of the tube adopted. The tube itself is shown in Figure 9. It consists of a flat molybdenum cathode between two forked anodes in a subminiature bulb. A connection is made by a clip and metallised strip to the inside wall of the bulb. Simplicity of design and low cost are likely to be a major consideration in a device of this type because of the large numbers needed in a practical system. The various tube parameters will now be described.

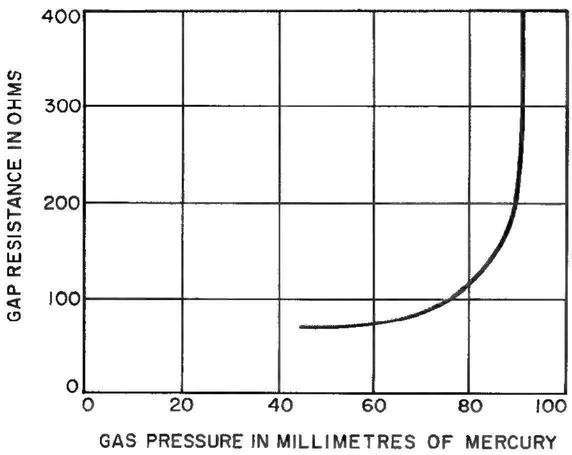


Figure 5—Variation of anode-to-anode resistance with helium gas pressure; anodes-to-cathode spacing is 2 millimetres (0·08 inch).

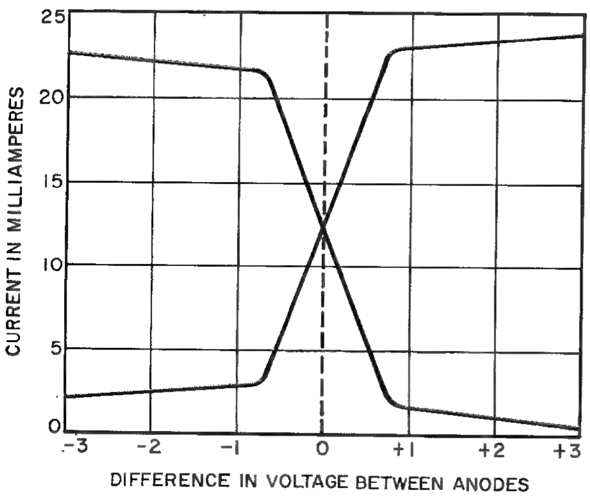


Figure 6—Static characteristics of speech gap. Resistance between each anode and the cathode is 1500 ohms; speech-gap resistance between anodes is 100 ohms.

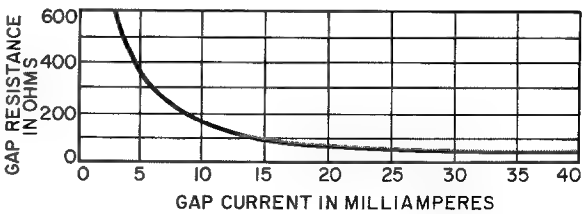


Figure 7—Variation of gap resistance with current.

3.1.1 *Switching Characteristics*

The requirements for use as a switch are:— Stable and reproducible static characteristics, minimum excess voltage for pulse breakdown, and, also, a minimum reduction in breakdown immediately after a conducting gap is extinguished.

3.1.2 *Maintaining Voltage V_m*

Because of the long-term stability required, it was decided to adopt the sputtered molybdenum technique, which was known from other work to be satisfactory. By this process, the cathode is cleaned by high-density ion bombardment, the sputtered material being deposited on the wall of the tube and thus preventing release of impurities from the glass during life. A connection is made to this sputtered layer to control wall charges that otherwise would cause variations in tube characteristics. By this means, stable 120-volt values of V_m were obtained.

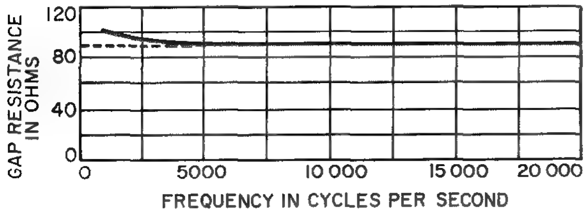


Figure 8 -Variation of gap resistance with frequency.

3.1.3 *Breakdown Voltage V_b*

The aim in this case is to obtain the maximum value consistent with low resistance; consequently, the pressure times distance product was fixed just below the knee of the graph of Figure 5, giving a V_b of 270 volts.

3.1.4 *Pulsed Breakdown*

It is well known that when a limited time is allowed for breakdown, the breakdown becomes subject to statistical variation and, furthermore, in the absence of these variations a glow formation time is observed that is related to the dark current and the excess applied potential. The former effect is eliminated and the latter reduced by priming. The possibility of using radioactive priming was considered and, in particular, the use of tritium was investigated. It appeared very attractive from many aspects, especially the low particle energy ($\approx 10^4$ electron–volts) and half-life of about 12 years. Tests showed tritium to be reasonably effective but difficuties were encountered in admitting known quantities into the tube and effectively controlling the process.

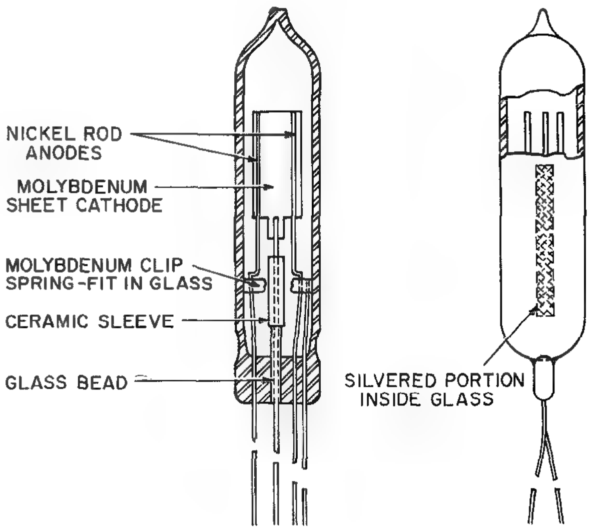


Figure 9—Construction of gas-tube switch.

Tritium was eventually replaced by an auxiliary glow discharge. This was arranged by dual use of the internal clip connection—which was connected to a negative voltage through a high resistance, A subnormal glow discharge of about 10^{-6} ampere was maintained to a selected point on the clip lead. This was found to be extremely efficient in priming the main discharge, giving pulse breakdowns with only a small overvoltage.

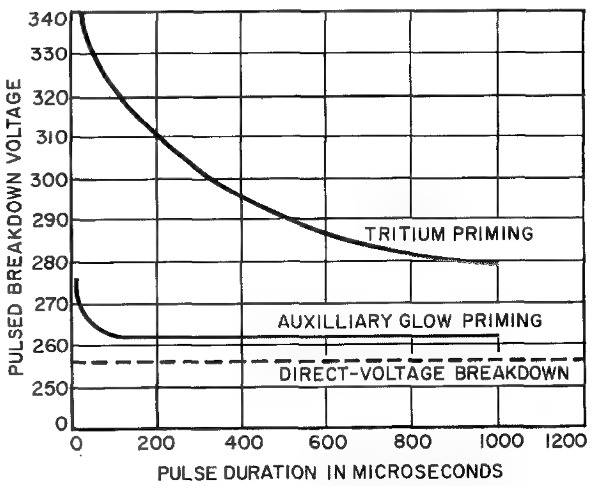


Figure 10—Pulsed breakdown voltage versus pulse width.

Dynamic breakdown characteristics are shown in Figure 10 for tritium and auxiliary primed tubes.

3.1.5 *Reapplied Anode Voltage*

Reapplied anode voltage is the highest voltage that can be reapplied to the gap after a specified time without causing a further breakdown. It is related to static breakdown by deionisation time and other residual effects. In the present tubes the time allowed is an order of magnitude greater than the deionisation time and the value obtained is largely controlled by cathode surface effects and depends on the magnitude of the current before extinction. The reapplied voltage may be 15 volts below the static breakdown voltage.

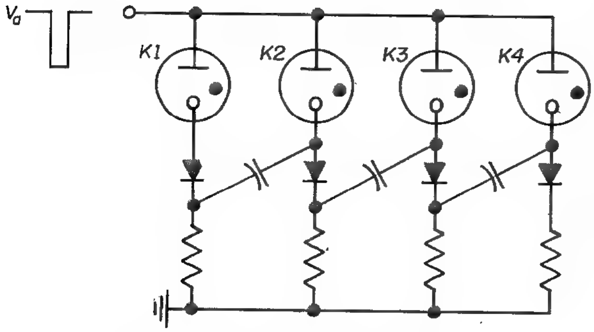


Figure 11—Diode switching circuit.

It is important to note that, though static breakdown can be controlled to within close limits, the dynamic limits are much wider. This is particularly true in the case of coordinate switching using double pulses superimposed on standing bias, since a single pulse applied immediately after extinguishing a gap may cause it to refire. Furthermore, standing voltages must be appreciably higher than the tube drop to permit sufficient current stabilisation, which is important in balancing the system. The characteristics obtained from a tube of this design provided satisfactory operating conditions for model exchanges.

Early life tests showed that characteristics changed and it was established that these changes were a direct result of the method of processing the tube. To minimise overall pumping time, the tubes were processed in neon to take advantage of the high sputtering rate. The neon was then pumped away and the final gas filling of helium admitted. It was found that neon was released from the cathode during life and causing the change in characteristics. Methods were subsequently developed for processing and filling only with helium as this was the most-suitable gas filling for the required characteristics. Satisfactory life was then obtained. In the application envisaged for this device, 4000-hour life with continuous conduction is equivalent to 25 years of normal service.

4. *High-Speed Triode-Switch Operation*

Attention was now directed towards the possibility of higher-speed operation. Initial studies showed that individual elements with external coupling showed more promise than multi-electrode tubes operating on the ionisation glow transfer principle. A new technique that largely removed the circuit limitation was evolved and permitted the use of tubes to the limit of their characteristics.[5] The basic principle is to extinguish a conducting gap by a negative pulse on the commoned anodes and to apply the resulting change in cathode voltage to initiate a discharge in another gap.

4.1 Circuit Description

This technique is suitable for either diode or triode elements. It was established, using *G1/371K* trigger tubes, that operation was feasible at input frequencies in excess of 100 kilocycles —limited by the deionisation time of 10 microseconds at low current level. The simplest case, using diodes, is shown in Figure 11. With anode voltage V_a set below breakdown voltage V_b, an output voltage $V_o = V_a - V_m$ appears on conducting cathode *K2*. This is the steady-state condition in which *K3* is restored to earth through the low forward resistance of the crystal diode. Application of a negative impulse voltage greater than $V_a - V_m$ extinguishes the glow on *K2* and the output voltage decays towards earth. This results in a similar change in voltage appearing at *K3*, negative with respect to earth and developed across the high reverse resistance of the crystal diode. Providing the sum of the negative change on *K3* and V_a is greater than the instantaneous V_b of gap *3*, discharge will be initiated on *K3*. For a given deionisation time, the maximum frequency is doubled by use of biphase

[5] British patent 33115/54.

waveforms connected to alternate anodes. The maximum operating frequency then occurs when the pulse spacing equals the deionisation time.

It was concluded that provided deionisation time, glow formation delay time, and build-up time could be substantially reduced, an element capable of working in the 0·5-to-1-megacycle range could be made. Independent investigations of the individual parameters involved were made.

4.2 Glow Initiation

Published work shows that short time lags for spark breakdown are obtained with low overvoltage (10^{-6} second with 1-per-cent overvoltage for 1-centimetre gap). These workers have invariably used high values of pressure times distance, greater than 200 millimetres of mercury times centimetres; a region where considerable controversy still exists regarding the exact mechanism of breakdown. The present experiments have been restricted to low products, approximately 20 millimetres of mercury times centimetres, where the validity of the Townsend model is undisputed. From earlier work, some difficulty was anticipated in obtaining a substantial reduction in glow formation time. Here this time is defined as the interval between application of a specified overvoltage and the establishment of a self-sustained glow. This requires field distortion by the formation of a positive-ion space charge in front of the cathode. For uniform fields, increase in priming current reduces the dynamic breakdown potential but does not greatly alter the overvoltage for a given time delay because the static breakdown voltage is also lowered. Only when priming is sufficient to reduce V_b to V_m is the formation delay eliminated. In this condition the space charge is already formed and occurs when a primed electrode is situated inside the cathode dark space of a priming discharge. Such conditions, however, are not useful when the breakdown is required to be considerably higher than the maintaining voltage. For this condition, therefore, with parallel plane electrodes, the field strength at the cathode is sufficient to produce breakdown at very-low values of ionisation current and short glow formation time can only be obtained with relatively large overvoltage. The obvious requirement is to provide a stable condition with a relatively high degree of ionisation.

Alternative geometries were examined and it was established that suitable conditions were provided by both concentric and spherical structures operating with positive wires and points. The field strength near the anode is greatly enhanced and the field at the cathode is reduced in the ratio of the radii. This results in high gas-current amplification and produces an ionised state in the anode region. The current density near the cathode, however, is comparatively low, resulting in a low secondary coefficient. The overall result is that high values of prebreakdown current are obtained without breakdown This stabilises the breakdown of the gap. If a concentric diode with a source of priming charges along its axis is considered, the high field strength in the anode region will produce ionisation there, resulting in the formation of a partial space charge and only a small amount of additional energy is required to cause breakdown. This results in a short glow-formation time. Typical curves are shown in Figure 12 for concentric and spherical gaps. Two methods have been devised for producing the required conditions.

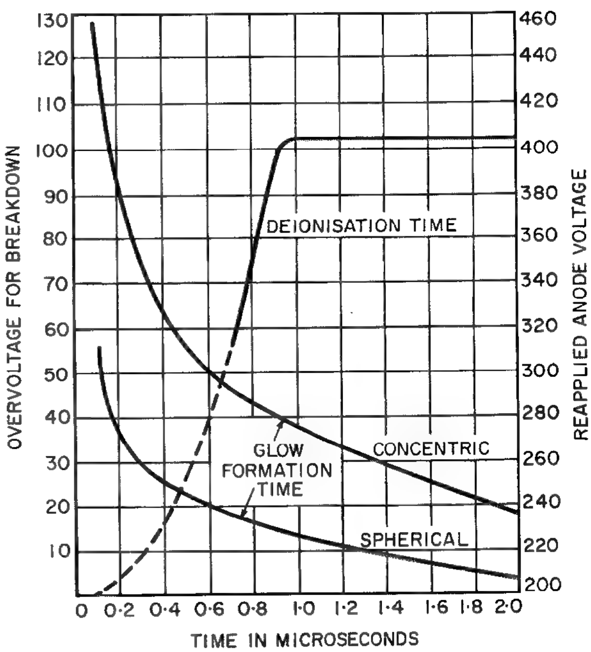


Figure 12—Typical breakdown and deionisation time characteristics for diode gaps. Priming current $\approx 10^{-8}$ ampere.

4.2.1 *Corona Discharge at Positive Wire*

A complete volt–ampere characteristic of a tube with a fine anode wire is shown in Figure 13, the initial electrons being obtained by using the anode wire as a source of light. The plot shows that the current increases with applied voltage up to point *B* where corona occurs and then increases sharply to point *C*, thereafter it increases at approximately 2 microamperes per volt to point *D*, where the gap breaks down into a self-sustained discharge. A similar effect is observed with anode cold, though at a higher anode voltage, depending on the wire radius and the dielectric strength of the gas. The important point here is that the formation delay times above and below point *B* are enormously different. With V_a set above *B*, delays are very short but below this point, time lags become very long. The successful application of corona priming would produce a very-simple and attractive device but some troubles have yet to be overcome. The main difficulty is the small voltage difference between corona breakdown and breakdown proper, a substantial differential being required to ensure fast-enough corona reformation time after application of an extinguishing pulse. This becomes important in application and involves the state of the next tube following a discharging tube. It will be remembered that the extinguishing pulse is applied in common to all anodes. If corona is extinguished in the tube concerned it must reform in a time less than the pulse width, otherwise the tube will be inadequately primed. The details are still being examined.

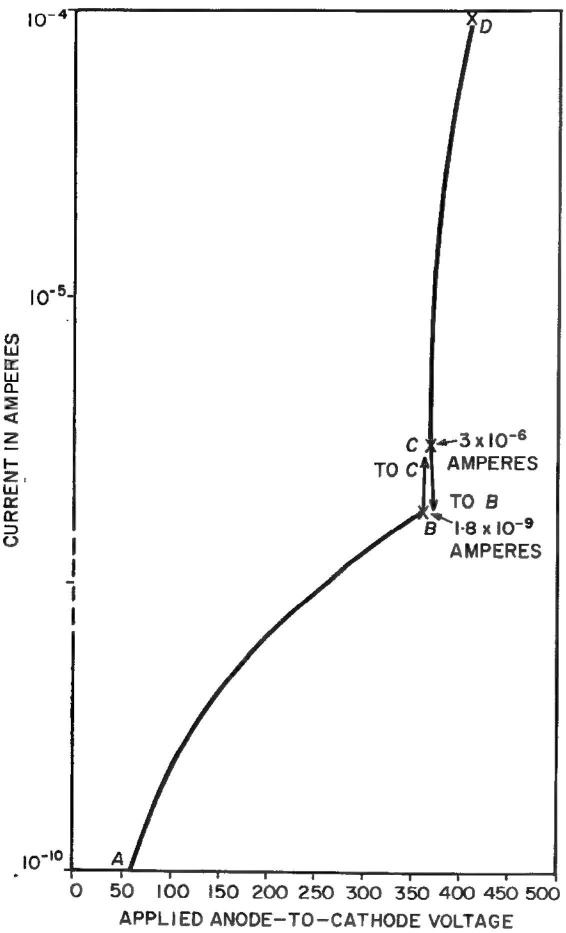


Figure 13—Voltage–current characteristic showing corona priming.

4.2.2 *Central Priming Discharge*

Similar stable prebreakdown currents are obtained in the constructions shown in Figure 14, which also shows a plot of the prebreakdown current against anode voltage.

4.2.2.1 *Concentric Geometry*

A fine tungsten wire is stretched along the axis of an apertured cylindrical anode inside a cylindrical cathode. By maintaining a direct-current discharge with the centre wire as cathode, some ionisation occurs in the main gap producing the breakdown characteristic plotted.

4.2.2.2 *Spherical Geometry*

Two fine wires are beaded on a ceramic disc which fits into a domed cup. This is a very-simple construction giving stable characteristics.

4.3 Deionisation

Considering first a discharging gap, the potential distribution is such that almost the entire applied voltage appears across the cathode dark space because the field is distorted by the presence of the positive-ion space charge. Within a certain specified time, the charge density must be reduced below a certain level, which, when the field is reapplied, will be insufficient to reform the discharge. Several factors involved in this process are listed below:

A. Gas filling and pressure (involving mobility K and pressure p).

B. Spacing d.

C. Sweeping field X.

D. Geometry.

E. Current.

F. Diffusion volume.

Before considering these points in detail, it should be mentioned that previous work had established the use of admixtures of hydrogen as a deionising agent. Metastable states of long lifetime are eliminated by the introduction of hydrogen and deionisation time can be reduced from milliseconds to microseconds. Any further reduction in time must be related to the mobility of the particles involved and, as electron mobility is at least 10^3 times greater than ion mobility, it is with the elimination of ions that we are concerned.

It has been found that the deionisation time τ is approximately equal to d/KX, the drift velocity being KX with $X = V/d$ being the field applied across the gap during the deionisation period and set just below the maintaining potential V_m. Voltage V_m is relatively independent of p and d over the range involved. Breakdown voltage V_b is set by circuit requirements and is a constant times p times d for a given gas.

$$\tau = d^2/KV_m$$

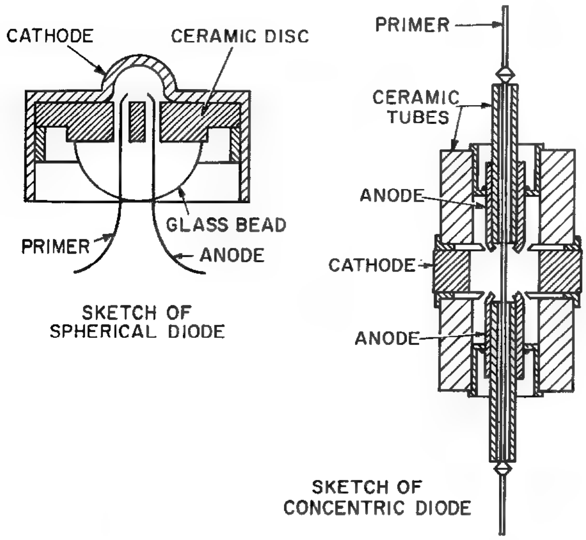
But $K = C_1/p$ and $p = C_2/d$. Thus,

$$K = dC_1/C_2$$

and

$$\tau \doteqdot dC_2/C_1V_m$$
$$\propto d.$$

In practice, as d is made smaller and p increased, the current density, which is inversely proportional to p^2, becomes very high resulting in contraction of glow and requires a small cathode area to maintain glow stability. Reduction of the cathode area magnifies edge effects and therefore conditions have to be optimised to the extent of obtaining usable values of deionisation time, V_m, output voltage, and current. The relevant factors controlling the selection of gas filling are positive ion mobility and dielectric strength, the latter permitting a reduced gap for a given breakdown. Hydrogen best satisfies these requirements but has a maintaining potential of about 300 volts, which involves high operating voltages. Mixtures of inert gases have therefore



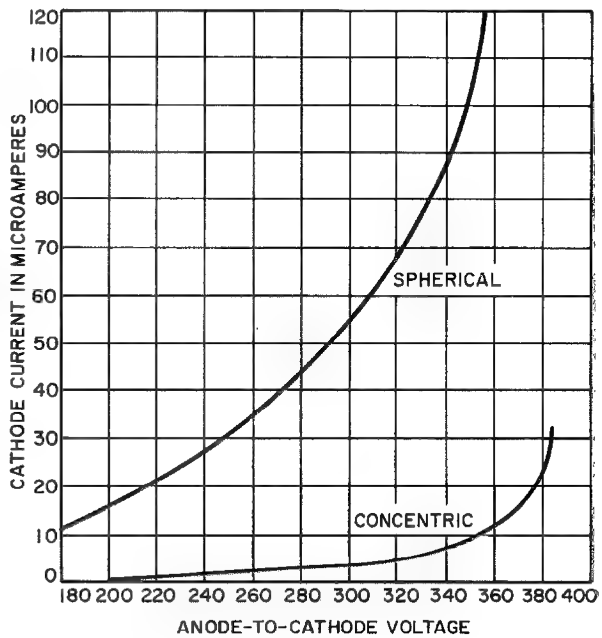


Figure 14—Prebreakdown current versus gap voltage for two tube structures. Priming current $\approx 10^{-8}$ ampere. Sketches are about 2·5-times actual size.

been used in which the hydrogen content is optimised to be most suitable for producing the required operating conditions.

4.4 Geometry

Most early work was carried out using parallel plate electrodes but it was apparent that the field strength at the cathode would influence the reapplied voltage level; a lower field would permit a larger amount of residual ionisation without refiring. Fortunately, therefore, structures that have been shown suitable for rapid glow initiation are also most suitable for fast deionisation time. In a conducting gap, the nonuniform field goes over to roughly parallel-plane conditions due to the presence of the positive ion space charge. On extinction, the important process is the collapse of the space charge, the field thereafter reverting to radial form. The net result is that a higher reapplied voltage is permissible after the collapse time, which can be interpreted as a shorter time for a reapplication of a given percentage of V_b. It is not claimed that geometry is the major factor in the reduced deionisation times (factor of about 100 overall), but rather the general scaling down. It is important to note, however, that the deionisation time becomes very long if the gap is reversed by using $-V_e$ wire. A typical deionisation time characteristic is shown in Figure 10 and may be compared with the corresponding curves for the *G1/371K*, shown in Figure 4, which is, at present, the highest-speed cold-cathode trigger tube available.

Tubes of the above designs have been demonstrated to operate at input recurrence frequencies up to 1 megacycle, but their final design has not yet been fixed.

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Transition-Rate Discriminators

By J. D. HOLLAND

Standard Telephones and Cables Limited; London, England

TELEGRAPH signalling waveforms consist of a series of mark–space or space–mark transitions that represent points at which the rate of change of information is a maximum. The number of transitions that occur over a given period for an error-free message depends on the signalling speed. If the message is disturbed by noise the transition rate increases, and for certain types of interference the rate may tend to become zero.

The difference in transition rate between an error-free message and one perturbed by noise can be detected, and the information used to provide a utilization voltage. Circuits have been developed for this purpose, known as transition-rate discriminators; they have a number of applications.

1. *Transition Rate of a Telegraph Signal*

Examination of the 5-unit start–stop telegraph code shows that 6 characters have 6 mark–space or space–mark transitions, 6 characters have 2 transitions, and the rest have but 4 transitions of either type. In contrast, the 7-unit error-detecting code has between 1 and up to 6 transitions per character.

With the 5-unit code the transition arrangement is shown in Table 1.

TABLE 1

5-UNIT-CODE TRANSITION ARRANGEMENT

Number of Characters	Number of Transitions	Characters
6	2	M, O, T, V, LETTERS key, and BLANK
6	6	D, F, J, R, S, and Y
20	4	All the rest

The transition rate can conveniently be expressed in terms of frequency as:

$$F = N/2D$$

where

F = transition rate in cycles per second

N = number of mark–space or space–mark transitions per character

D = duration time of a character in seconds.

Hence, for the 5-unit code at 50 bauds the rate for 2-, 4-, or 6-transition characters is closely equal to 6·6, 13·3, and 20 cycles per second, and the average number of transitions per character taken over a large number of randomly chosen characters is 4, that is, at a rate of about 13·3 cycles per second.

For example, in the standard test sentence, "The quick brown fox jumped over the lazy dog's black," the average rate is about 14·3 cycles per second.

An analysis of encoded or clear transmissions or a mixture of both infers that, at 50 bauds, the minimum and maximum rates are closely 10 and 18 cycles per second respectively. This does not include reversals (RYRY ...) that provide no semantic information and are discarded. (The rate for this signal type is about 21 cycles per second.)

2. *Frequency of Fluctuation Noise in a Given Bandwidth*

Rice[1] has shown the noise currents in a narrow-band filter behave like a sine wave of frequency $(f_1 + f_2)/2$ and of an amplitude that fluctuates at an irregular frequency of $(f_1 - f_2)/2$, where f_1 and f_2 represent the band limits of the filter. The output from the filter therefore resembles a single frequency, equal to the midband frequency, modulated in amplitude and phase. Since the amplitude modulation frequency approaches one half the filter bandwidth it seems reasonable to suppose that the phase modulation frequency does also.

If this spectrum is limited, applied to a demodulator and then to a frequency counter, the counter will register one pulse per cycle and the noise count will be closely equal to one half the bandwidth in cycles per second.

[1] S. O. Rice, "Mathematical Analysis of Random Noise," *Bell System Technical Journal*, volumes 23 and 24, numbers 1 and 3; July, 1944 and January, 1945.

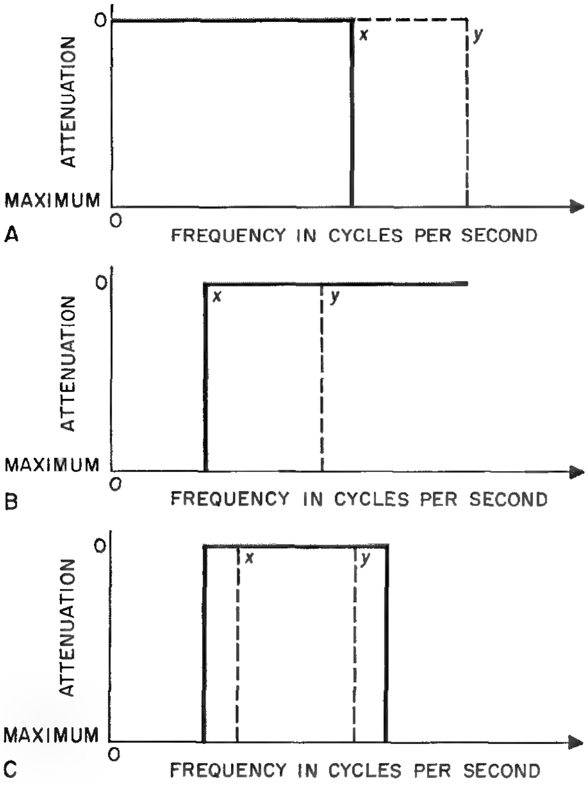

Figure 1—Characteristics of transition-rate discriminator.

3. Discrimination Between Noise and Signal Message Frequencies

The bandwidth in cycles per second required before demodulation of a telegraph signal of quasirectangular form is greater than the message transition rate in cycles per second. For example,[2] at 50 bauds, a bandwidth of about 100 cycles per second is required and hence the noise count will be about 50 cycles per second whereas the maximum rate due to a meaningful message is closely 18 cycles per second.

This difference in frequency can be detected by a transition-rate discriminator, and the type of action obtained is shown in Figure 1A, 1B, and 1C. In Figure 1A, at x, maximum attenuation is obtained to frequencies exceeding those that could occur in an error-free message and point y is another arbitrary cut-off point that would be used at higher keying frequencies; necessitating increased bandwidth.

In Figure 1B, maximum attenuation is obtained to frequencies lower than the lowest message frequency at the arbitrary cut-off points x

[2] L. J. Heaton-Armstrong and J. D. Holland, "Direct-Printing Receiving Systems at Low Radio Frequencies," *Electrical Communication*, volume 35, number 3, pages 202–208; 1958.

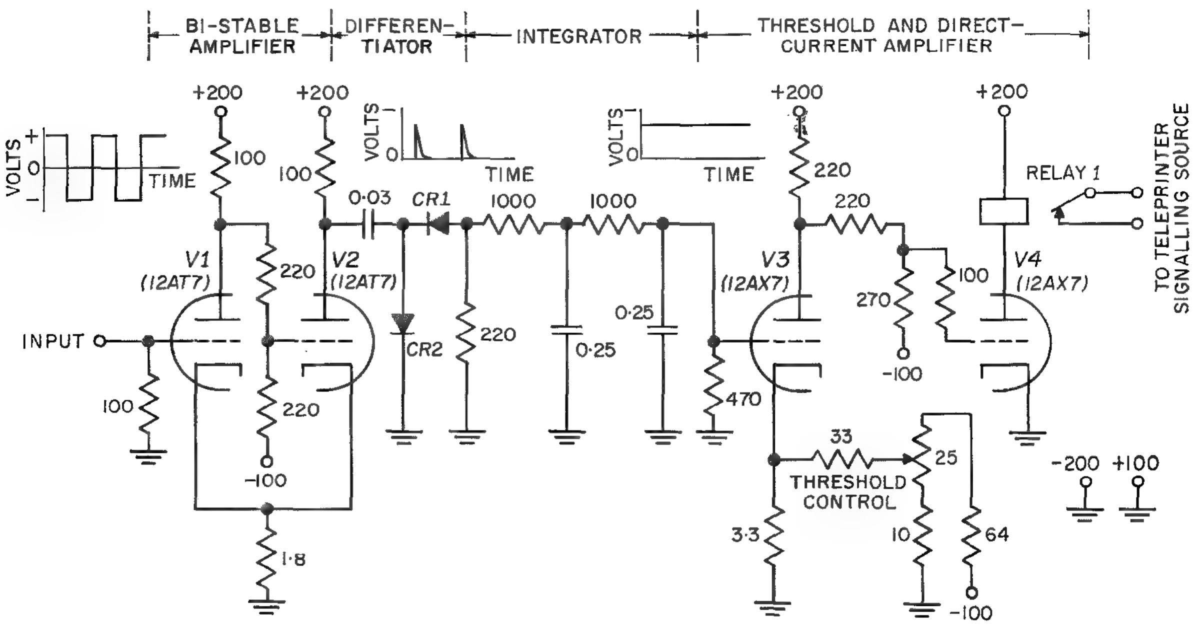

Figure 2—Transition-rate discriminator for 50-baud operation. Values of resistance are in kilohms and values of capacitance are in microfarads.

and y. Figure 1C shows the type of characteristic that can be obtained by the use of two discriminators of the characteristics shown in Figure 1A and 1B. In this case, frequencies above or below the band occupied by an error-free message are suppressed.

4. *Transition-Rate Discriminators*

One form of circuit[3] and the functions obtained is shown in Figure 2. Due to the use of a bistable amplifier the positive-going output from *V2* can be short-circuited by use of diode *CR2* since the transition history of either polarity of the wave is identical.

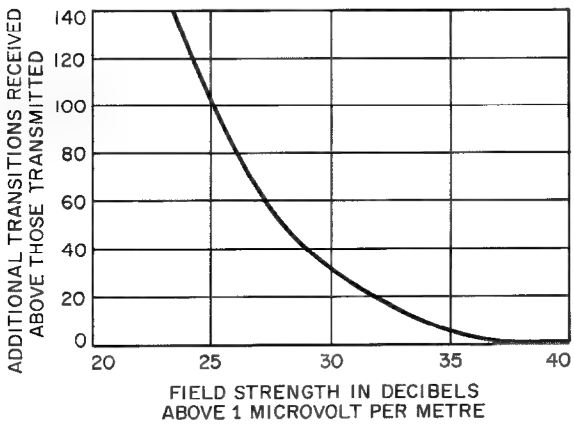

Figure 3—Spurious extra transitions received as a function of the field strength of the received radio signal.

This form of circuit is suitable for connection to the output of a demodulator. For example, the number of extra transitions received above those originally sent in a random message against field strength is shown in Figure 3. These characteristics were taken at the output of the demodulator of an *RV.14* receiver. Error-counting tests showed that the error rate was about 1 in 1000 at a field strength level of 30·3 decibels relative to 1 microvolt per metre.

Figure 4 shows the results obtained after demodulation, using reversal-type signals or signals of random character from a high-frequency receiver. The signal-to-noise ratio of the receiver was approximately 12 decibels for an input of 0·1 microvolt in 75 ohms.

Valves *V1* and *V2* (Figure 3) can be dispensed with and connection made to the output terminals of the telegraph receiver, providing the in-built circuits are bi-stable. Similarly valves *V3* and *V4* can be omitted if the negative-going utilization voltage across the 470-kilohm resistance is required for other applications. Some form of threshold circuit should be retained.

For the circuit shown in Figure 2, the characteristic of the output level versus frequency across the 470-kilohm resistor is shown in Figure

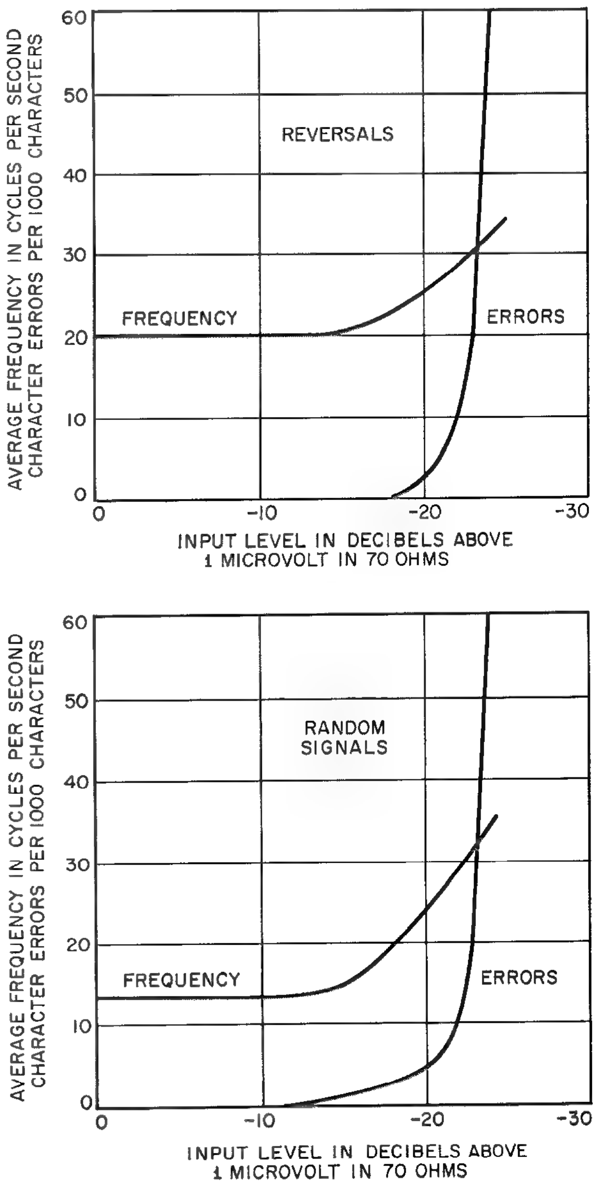

Figure 4—Average message frequency in cycles per second or character errors per 1000 characters as a function of input signal level for two types of signal. Signalling speed = 50 bauds; bandwidth = 120 cycles per second.

[3] British Patent Application 7077/57.

5 and the input required at the grid of $V1$ should not be less than ± 5 volts. The threshold control is effective to within a change of ± 1 cycle per second of the transition frequency.

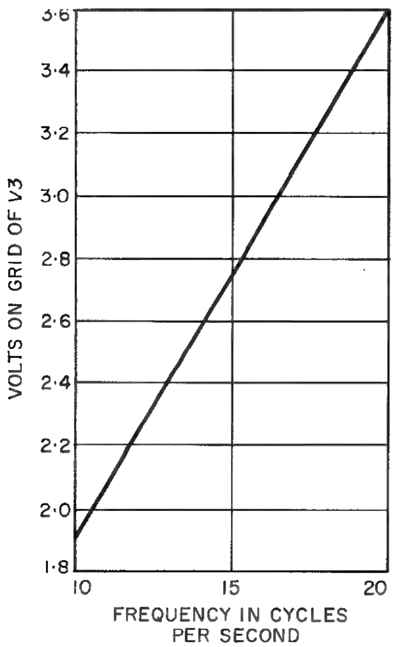

Figure 5—Voltage appearing on the grid of $V3$ as a function of frequency with an input of not less than +5 and −5 volts on the grid of $V1$.

5. *Application*

5.1 Prevention of Printing Random Symbols

At 50 bauds, the threshold control of the circuit shown in Figure 2 is set so that the relay contacts are opened at 19 cycles per second.

If the keying speed is not known, correct operation is obtained by monitoring on a reversal signal (RYRYRY...) and adjusting the threshold control until repetition of these characters is eliminated.

5.2 Signal Combination

When signalling currents carrying the same information, but arriving over different channels or radio paths, are combined, the effective output after combination will be mutilated if any of the sources are perturbed by noise of a higher level than the other contributions. This has been checked in the field on a dual diversity system. It was found that without transition-rate discriminators in each channel the output, with severe noise from one path, was completely mutilated, whereas with the discriminators in circuit before the combining process, the error rate was of the order of 1 in 2000 characters.

5.3 Dual-Diversity Switched-Antenna Receiving System

Dual-diversity spaced-antenna systems are of necessity complex in that separate high-grade receivers are used requiring periodic alignment checks to verify that equal weighting is being given to each path over the frequency range.

To reduce equipment investment, systems have been proposed whereby one receiver is switched between antennas in such a manner that the receiver is always connected to the antenna offering the highest signal level.

Tests show[4] that the error rate is reduced as compared to reception on a single antenna, and although this rate can never be as low as a system employing separate receivers and proper combining methods, a considerable saving in equipment is obtained.

The difficulty with the switched-antenna scheme is that since the switching information is obtained on an amplitude basis, the action required is likely to be indecisive when high noise conditions exist at one antenna site and a usable signal exists at the second antenna.

A system is proposed[5] in which the switching information is obtained from a transition-rate discriminator, having characteristics as shown in

[4] "Report on the Testing of an Antenna Diversity Equipment," internal report of Standard Elektrik Lorenz A. G., Pforzheim; 1956.
[5] British Patent Application 24649/58.

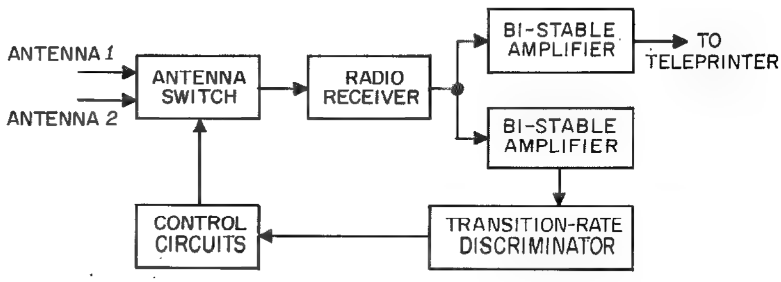

Figure 6—Diversity system using antenna switching to select best signal.

Figure 1A, in such a manner that the receiver is switched to a second antenna before reception on the first antenna becomes poor enough to cause appreciable errors.

A block diagram of the system is shown in Figure 6, and the detailed connection of the control circuits are shown in the reference.

The demodulated signal from the receiver goes to two bi-stable amplifiers. The triggering level of the amplifier actuating the teleprinter is so arranged that it will change state close to the steady-state condition of the wave whereas the amplifier forming part of the transition-rate discriminator changes state close to the zero level datum of the demodulated wave.

It can be shown[6] that, if the triggering level is the same as the steady-state condition of the wave, the value of the signal-to-noise ratio before a mark is turned into a pseudo space is −6 decibels; that is, the noise must exceed twice the momentary signal level. Alternatively, the ratio approaches unity for triggering levels close to the zero voltage datum line; and therefore the decision to switch should occur before a mutilization is recorded, provided that the fading rate is not too rapid.

[6] D. A. H. Johnson, "Laboratory Report 10," New Zealand Post, Telegraph, and Telephone Administration; 1957.

6. *Summary*

The probability of the presence of a signal in noise depends on the degree of *a-priori* knowledge of its precise form. Usually the maximum amplitude or the total received energy is examined and an existence probability is determined. Transition counting may represent a useful aid, since additional information, in terms of transition rate, is derived from the waveforms apart from an examination of their amplitude or energy content.

Probability Theory in Telephone Transmission*

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TELEPHONE switching problems mean endeavour to get a call connected through a desired channel with a certain specified expectation of success.

In long-distance telephone transmission the endeavour is to transport speech information over long distances with a specified expectation of the impairment that the speech signal will suffer.

A common endeavour in these two fields is to carry out the separate purposes in the most economical way and in doing this it is usually necessary fully to exploit "the specified expectation" since a design that gives too good a performance will usually be more expensive than one that only just meets the requirements.

One very important cause of impairment in speech transmission is noise. In long-distance telephone circuits there are several mechanisms that produce noise. The most important ones are:—

(A) Thermal agitation.

(B) Nonlinear (intermodulation) distortion in multichannel amplifiers.

(C) Disturbances from other circuits.

The magnitude of noise in the circuit due to these mechanisms can generally speaking be reduced to almost any desired degree but such reduction strongly affects the cost of the equipment required and it is therefore very important to find out how much noise can be tolerated. This meets with two difficulties:—

(A) The tolerable noise power at any time depends on the volume of the speech signal that is available at the end of the circuit at that time and this volume varies from call to call due to many factors, see for instance reference 2, page 29.

(B) The amount of noise present in the long-distance circuit will fluctuate with time, for instance, that part of the noise that arises from nonlinear distortion in a multichannel carrier system will fluctuate with the signal loading of the amplifiers[1–3] and in a radio link the thermal agitation noise will fluctuate due to fading of the radio path. The fluctuation involved is that of the mean power taken over a few hundred milliseconds and not the faster fluctuation that is present in thermal agitation noise.

One way of specifying the acceptable amount of noise is shown in Figure 1A. This specification amounts to allowing a certain maximum value (−50 decibels relative to 1 milliwatt) of noise power at a particular point in the long-distance circuit but with the escape clause that for 1 per cent of the time (in general during the busy hour when high noise is expected) no limit is placed on the amount of noise.

This way of specifying the noise is very simple but it does not take into account the full facts of the actual situation. It is not a very complete statement of the amount of noise that can be tolerated without seriously disturbing telephone connections. If in a particular design the noise statistic should be a straight line with a certain slope, then, for maximum noise of this type, the noise statistic line would have to meet the specification at the 1-per-cent point; and everywhere else the noise would be much smaller than would be acceptable.

This is an unnatural restriction and the result of the simple way in which the specification in Figure 1A is framed.

In Figure 1B is shown a more-recent proposal for a specification for the permissible noise, which is now limited in two ways.[5]

* Reprinted from *Teleteknik* (Copenhagen), volume 1, number 1; 1957. Presented at First International Congress on Application of the Theory of Probability in Telephone Engineering and Administration; Copenhagen, Denmark; June 20–23, 1955. The purpose of this paper is to point out that probability theory is used not only in telephone switching but also in telephone transmission. This is done by giving an example.

[1] All references will be found in the bibliography.

(A) The mean power of the noise over the whole hour shall not exceed -50 decibels relative to 1 milliwatt.

(B) The statistical distribution of the noise shall fall below the line shown.

This specification has been derived by taking account as far as possible of the known fact that the speech volume will fluctuate from call to call and the likelihood that the noise in actual systems will also fluctuate. The sloping line in Figure 1B is part of a log-normal distribution curve; that is, one in which the logarithm of the variant is normally distributed. This type of distribution is very appropriate for telephone transmission and is followed fairly accurately by a number of variables. The speech volume at a certain point in the telephone network has been found to be distributed in this way with a standard deviation of the order of 6 decibels. Now Figure 1B is in some cases easier to satisfy than Figure 1A but at the same time the actual nuisance value of the noise is no greater than for the specification in Figure 1A. Figure 1B is a more-accurate statement of the noise that actually can be tolerated. While Figure 1A might be called a convention, Figure 1B takes more account of the actual situation. The specification in Figure 1B has the advantage that families of analytical curves can be derived from it, each of which meet the overall specification (Figure 1B) in two points—they have the correct total mean power over the hour (average power *not* the logarithm of power (decibels)) and the extreme peaks meet the peak specification at 0·03 per cent. In Figure 1C is shown two extreme curves *a* and *c* which meet the specification—one suitable when the actual noise is very constant (*a*), and the other (*c*) suitable when the noise is highly fluctuating. A third curve *b* is shown as a representative of a whole family of similar curves that may be drawn—each with the required mean power over the test hour and all meeting the peak part of the specification in Figure 1B at its highest point. This particular family of curves has been named "augmented log-normal distribution curves"; they are in fact log-normal curves to which a constant power has been added. The mean power of the log-normal curve itself is in logarithmic measure: Median $+ 0{\cdot}115\,\sigma^2$ where σ is the standard deviation in decibels.

The sum of the mean power of the log-normal distribution and the added constant power is kept constant at the specified value for the family of curves.

Both the statements Figure 1A and Figure 1B are intended to apply to circuits of lengths of 2500 kilometres and now comes a problem: how are we to specify the requirements for a shorter length of circuit (partial circuit) such that when a number of partial circuits are put together in tandem to produce a circuit 2500 kilometres long the resultant noise will meet the specification.

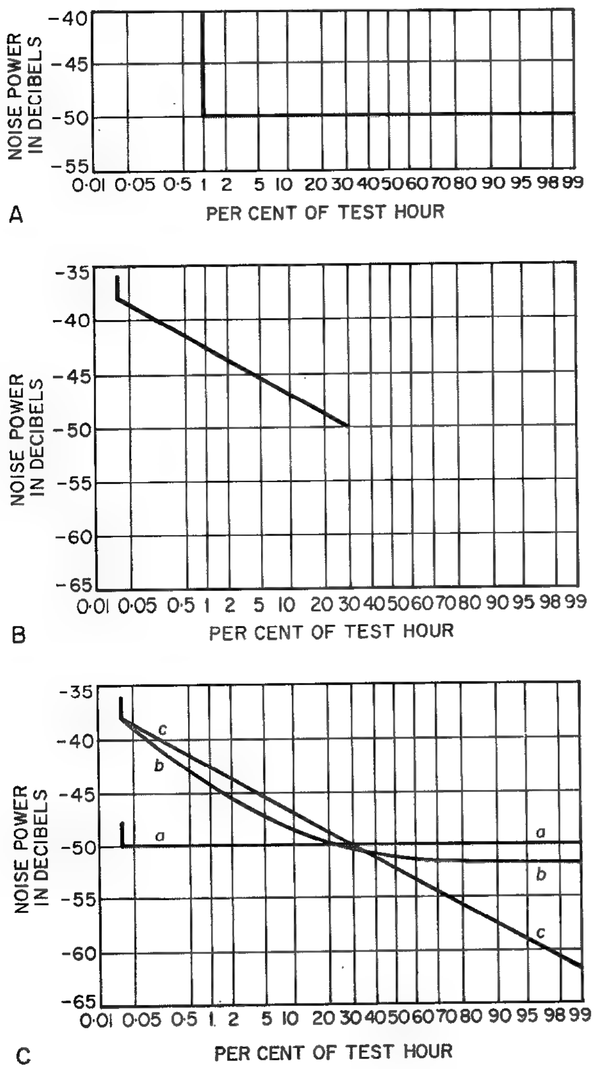

Figure 1—Noise specifications for a long-distance circuit of 2500 kilometres.

This is a difficult statistical problem and work is still proceeding[6] but a little can be said about two parts of the specification for a partial circuit.

(A) For the part of the curve that restricts the incidence of very-loud noise it is necessary that the frequency of occurrence in a partial circuit shall be proportionately smaller than for the full circuit, but in testing a single partial circuit, the test period may be lengthened in inverse proportion to the relative length of the partial circuit (see appendix 1).

(B) For the mean power over the test hour it is necessary that in the partial circuit this should be proportionately smaller than in the overall circuit but here it is also permissible in the case of the partial circuit to judge the mean noise power over a proportionately longer time. It would, of course, still have to be judged during times when a high noise level would be expected to occur. (This assumes that the busy hour et cetera will be the same for all regions through which the circuit passes. This will not always be true but is a safe assumption.)

It is however much-more difficult to find a proper answer for the main part of the curve which is not covered by (A) and (B). What we would like is to be able to specify the statistical properties of the noise that could be tolerated in for instance 1/9th of the total circuit which is specified in Figures 1B and 1C. The specification should be such that when 9 equipments having this performance are connected together the overall noise should just satisfy the specification in Figures 1B and 1C.

In combining the noise produced in the separate partial circuits it should be borne in mind that the variant to be summed up is power —not the logarithm of power (decibels).

For the present however it is probably necessary to "reverse" the problem and to propose a noise statistic for the partial circuit and then by computation find the noise performance that would be obtained if such partial circuits were connected in tandem to the length of 2500 kilometres and check if this meets the overall target.

1. Appendix

The lengthening of the test period for a partial circuit is justified by the following considerations.

The performance of the full circuit for a test hour may be considered to be the outcome of a repeated sampling process. The size of the sample must be proportional to the number of (equal) partial circuits required to make up the full circuit.

The population that is sampled contains the "hourly performance of the partial circuits" (for a period of the day when high noise is expected).

When only a single partial circuit is available and it is permissible to assume that further partial circuits will be similar, than the requisite size of sample may be taken from the hourly performance population of this circuit, or in other words, the test period for a partial circuit should be proportionally longer than the test period for a full circuit.

The individuals in a sample should be combined as if each represented the performance of one of the transmission sections that together make up a full circuit.

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Wide-Band Ultra-High-Frequency Over-the-Horizon Equipment*

By R. A. FELSENHELD, H. HAVSTAD, J. L. JATLOW,
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WIDE-BAND over-the-horizon radio equipment in the 680-to-900-megacycle-per-second band suitable for use in toll-quality multichannel telephone or television circuits is described. The system comprises 60-foot (18.3-meter) parabolic antennas fed with dual polarization horns, 10-watt drivers, 10-kilowatt amplifiers using 6-cavity klystrons, and receivers that permit dual or quadruple diversity by combining the received signals at the intermediate frequency. The over-all 1-decibel bandwidth is 15 megacycles, and the time delay distortion characteristics are suitable for interconnection with existing toll-quality radio links.

• • •

The broad-band tropospheric communication system described can transmit between 120 and 600 telephone channels with trunk-circuit quality or a television program, up to at least 200 miles. The equipment also interconnects with common-carrier telephone networks.

For the 185-mile (298-kilometer) link between Florida and Cuba, with a path loss of 197 decibels, this equipment, employing space diversity, was expected to provide a mean carrier-to-noise ratio of about 42 decibels and a received-signal level above the frequency-modulation threshold at least 99.9 percent of the time. With 120 telephone channels using frequency-division multiplex, the noise on each channel was expected to be less than 26 decibels adjusted at the −9-decibels-below-1-milliwatt point 99.9 percent of the time. Actual measurements showed the channel noise level to be about 22 decibels adjusted at zero level. There were definite indications that the terminal equipment rather than the tropospheric scatter equipment contributed largely to the channel noise.

Quadruple diversity, improving performance by about 3 decibels, can be obtained by combining space and frequency diversities. This combination is possible if an active stand-by system is operated on separate carrier frequencies. The system can also be used as two dual-diversity systems, permitting preventive maintenance on one system while the other is in service. It is also possible to transmit messages and television simultaneously, each on a dual-diversity reception basis.

This over-the-horizon equipment is terminated at the 70-megacycle intermediate frequency, at which frequency the diversity combining takes place. It is thus possible to connect to either line-of-sight or over-the-horizon repeaters at intermediate instead of base-band frequency. Table 1

TABLE 1
SYSTEM SPECIFICATIONS

Features	Requirements
Transmitter Power	10 kilowatts
Antennas	60-foot (18.3-meter) paraboloidal reflectors (two at each end) with dual-polarization horns
Diversity	Quadruple (space and frequency); may be used as two dual diversity systems; intermediate-frequency (70-megacycle) combining used
Service	System input and output 70-megacycle frequency-modulation signal; capable of handling at least 120 telephone channels; can also be used simultaneously for television transmission
Frequency Band	680–900 megacycles
Transmission Lines	*WR-1150* waveguide for transmitting line, pressurized; $3\frac{1}{8}$-inch (7.94-centimeter) Styroflex cable for receiving line, pressurized
Bandwidth	15 megacycles, 1 decibel down; 19 megacycles 3 decibels down: with suitable terminal equipment, 6 megacycles, 1 decibel down, video-to-video

* Reprinted from *Communication and Electronics*, number 35, pages 86–93; March, 1958. Presented, American Institute of Electrical Engineers' Winter General Meeting; February 2–7, 1958.

TABLE 1—*Continued*

Features	Requirements
Envelope Delay Distortion	Suitable for multihop, multichannel telephone or wide-band television transmission
Noise Loading Distortion Products	More than 50 decibels below signal level
Safety Features	All equipment fully interlocked and protected; key interlocks used in power amplifier
Alarm and Metering Circuits	All equipment fully equipped with alarm circuits and contacts for external alarms; metering and test points incorporated for alignment and maintenance
Environmental	Continuous operation in tropical climates

lists the specifications for the system, and Figure 1 shows its major elements and the switching facility provided.

The first application of this broad-band equipment is the microwave communication system between Miami and Havana jointly sponsored by American Telephone and Telegraph Company and International Telephone and Telegraph Corporation. The system includes two *TD-2* line-of-sight hops in Florida between Miami and Florida City where it connects with the over-the-horizon hop to Guanabo, Cuba. A *TD-2* line-of-sight hop completes the system from Guanabo to Havana. This system supplements the present cable system between Florida and Havana to provide many more message channels. It also provides a two-way television channel between the United States and Cuba.

1. Converter–Amplifier

The converter-amplifier accepts a 70-megacycle frequency-modulation signal, converts it to frequencies in the 692-to-880-megacycle band and amplifies it to provide sufficient driving power for the 10-kilowatt klystron power amplifier. Specifications are as shown in Table 2.

A block diagram of the converter–amplifier is shown in Figure 2. Local-oscillator frequencies are selected in the range from 51.8 to 59.4 megacycles to minimize spurious responses in the 70-megacycle band in the intermediate-frequency

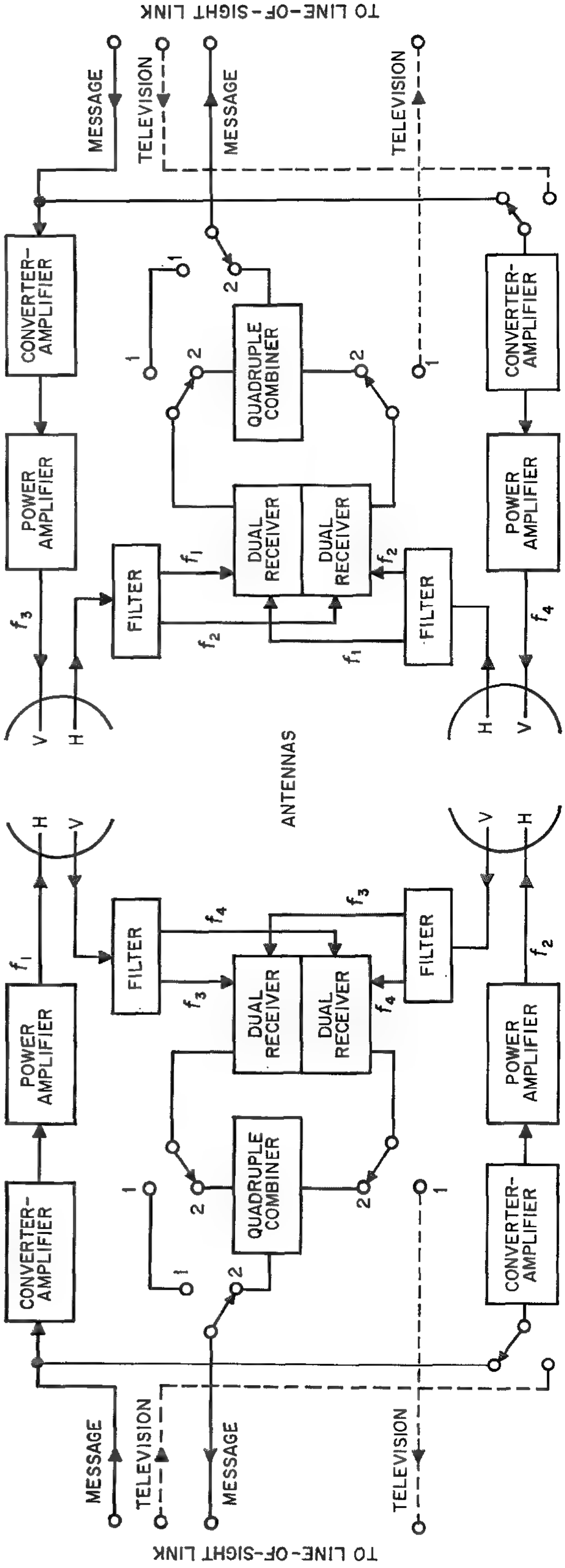

Figure 1—Block diagram of complete wide-band over-the-horizon system showing major components.

arm of the mixer. Local oscillator frequencies are multiplied and amplified by the multiplier chain and applied to the mixer at a fixed level of approximately 4 watts. The 70-megacycle input frequencies are amplified and applied to the mixer at an adjustable level of approximately 0.5 watt. The beat frequency obtained in the mixer is then selected by tuned circuits and amplified to the required power level.

The stability of the output frequency of this system is a function of the stability of the local oscillator and of the incoming 70-megacycle signal. To reduce the total carrier frequency drift to a value acceptable to the system, the local oscillator has been designed for a total frequency change of one part per million per day. The multiplied local-oscillator frequency is above that of the carrier for output frequencies of 840 and 880 megacycles and below that of the carrier for 692 and 740 megacycles.

The converter–amplifier consists of four 19-inch (48.2-centimeter), rack-mounted chassis in a 7-foot (2.1-meter) cabinet with front and rear access doors, as shown in Figure 3. The top half contains the radio-frequency chassis, with the

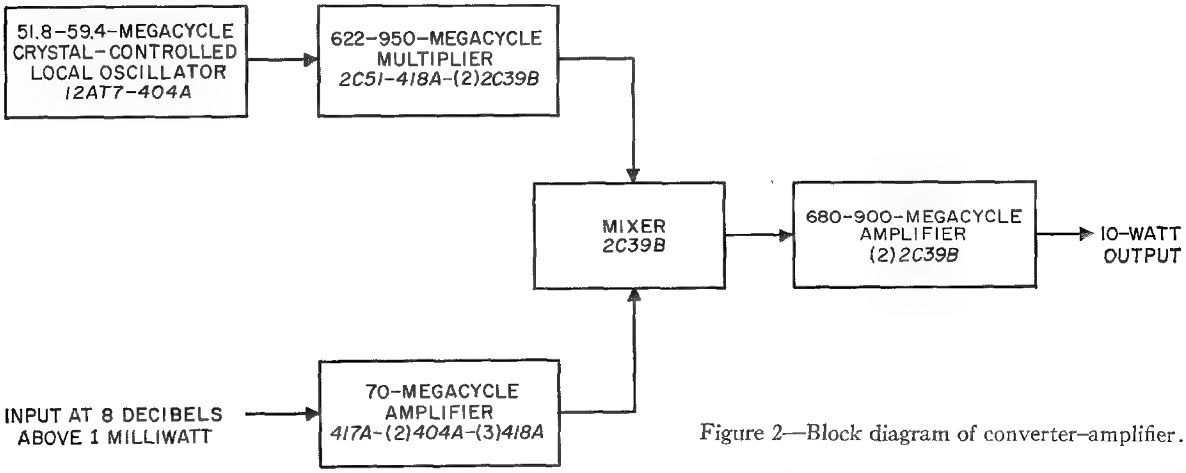

Figure 2—Block diagram of converter–amplifier.

TABLE 2

SPECIFICATIONS FOR CONVERTER–AMPLIFIER

Features	Requirements
INPUT	
Power	+8 decibels above 1 milliwatt minimum
Frequency	70 megacycles
Bandwidth	20 megacycles
Impedance	75 ohms
OUTPUT	
Power	10 watts
Frequency	680–900 megacycles
Bandwidth	20 megacycles at $\frac{1}{4}$-decibel points
Impedance	50 ohms
Envelope Delay	<4 nanoseconds* for ±4 megacycles

* Nanosecond = 10^{-9} second.

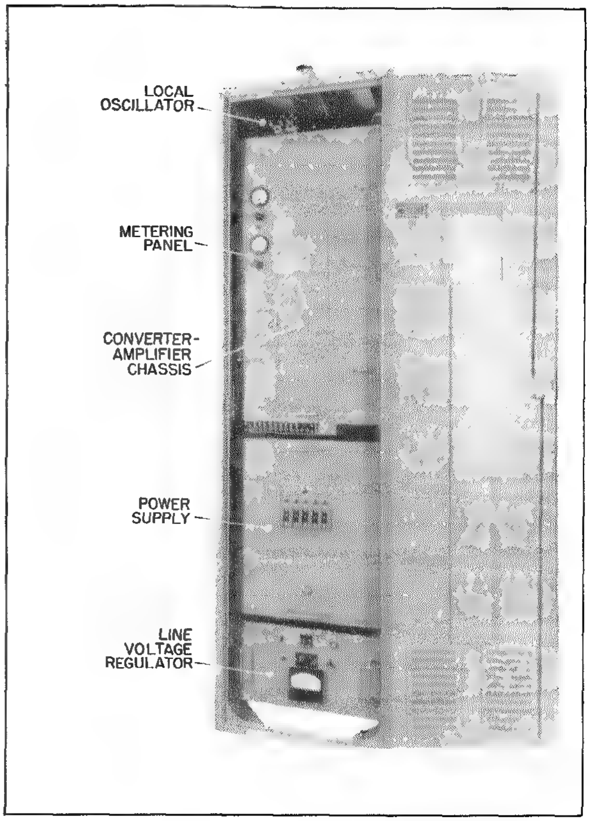

Figure 3—Front view of converter–amplifier.

power supply and voltage regulator in the lower half. The operating facilities on the front that are accessible without interrupting the interlocks are: the metering panel, tuning of local oscillator and multipliers, and tuning of mixer amplifier. The last two are available through doors in the protective dust cover. Power control and circuit breakers and a line-voltage meter are available on the front panel. A ventilating fan, with replaceable filter, is mounted on the inside of the rear door. The rear door is interlocked with the high-voltage-controlling relays and when opened the high voltage is turned off.

1.1 Radio-Frequency Chassis

The radio-frequency chassis, Figure 4, in addition to providing a mounting surface for the detachable intermediate-frequency power amplifier and radio-frequency multiplier subchassis, also contains the mixer-amplifier chain, metering facilities, monitoring circuits, fan and manifold, interlocks, and alarm circuits.

Three types of metering and monitoring circuits are available.

A. Two panel meters to monitor certain points that indicate satisfactory operation of the equipment.

B. Test points, on a jack panel accessible without turning the equipment off, with the proper test equipment to determine transmission characteristics of the equipment.

C. Each vacuum tube is connected to a test jack for measuring the cathode current by one of the panel meters.

The ventilation system is on the rear of the chassis and consists of a 115-cubic-foot- (3.3-cubic-meter-) per-minute blower with a distribution manifold and individual air ducts leading to each tube to be cooled. Failure of the blower is indicated by a relay and alarm buzzer activated by an air-flow switch, as well as by a lamp on top of the cabinet.

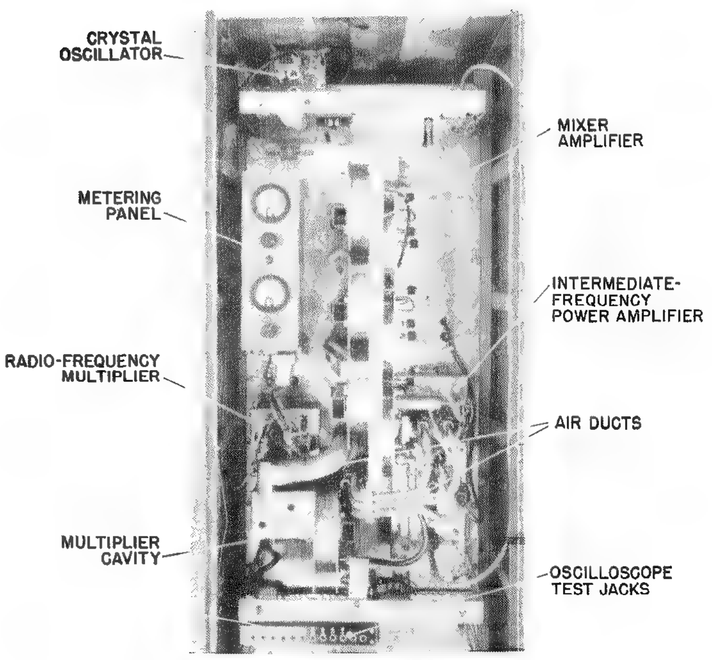

Figure 4—Detailed front view of radio-frequency chassis, converter–amplifier.

Two devices are used for protection against exposure to high voltage. The first device employs interlock switches in the front dust covers and rear cabinet access door. The second employs gravity-operated short-circuiting bars, operated by removal of the front dust covers and the rear access door.

1.2 Stable Oscillator

The oscillator consists of a stable crystal-controlled oscillator and a buffer amplifier. The tubes are one *12AT7* and one *404A*. The crystal is temperature-controlled and operates in its fifth overtone from 51.8 to 59.4 megacycles with a

stability of 1.0 part per million per day. The output is about 2 volts into a 90-ohm load.

1.3 Radio-Frequency Multiplier

A chain of four tubes multiplies the frequency received from the local oscillator and amplifies the power to a level sufficient to drive the high-level mixer. The multiplication factors are 12 for output frequencies of 622 to 670 megacycles and 16 for output frequencies of 910 to 950 megacycles. The tubes used are one *2C51/396A*, one *418A*, and two *2C39B*'s. The relative power levels and frequency distributions along the multiplier chain are shown in Figure 5. One volt across the 90-ohm input impedance is required for a 3-watt output.

The coupling circuits used between the first three tubes are conventional lumped-constant tuned circuits with a radial transmission-line cavity terminating the fourth tube. An adjustable coupling loop in this cavity couples the power to the input of the mixer.

The envelope temperature is kept below approximately 100 degrees centigrade by air cooling on all tubes except the *2C51* and the total power dissipation in any tube is below 50 percent of the rated dissipation for the tube.

Figure 5—Block diagram of radio-frequency multiplier in converter–amplifier. Total frequency multiplication is 12 for the low band and 16 for the high band. Oscillator frequencies are 51.833 to 55.833 megacycles for the low band and 56.875 to 59.375 megacycles for the high band.

1.4 Intermediate-Frequency Power Amplifier

The intermediate-frequency amplifier is a broad-band 6-tube 5-stage unit that amplifies the 70-megacycle signal. It will deliver an output power adjustable between 0.1 to 1.5 watts with an input signal of 7 milliwatts across 75 ohms. The tubes used are one *417A* grounded-grid input stage, two *404A*'s and one *418A* interstage, and two *418A*'s in a push-pull output stage; see Figure 6.

The input coupling network is provided with a matching adjustment consisting of bias control of the input tube and a single-tuned broad-band cathode inductance. A matching adjustment is necessary because the input to this amplifier may be from a relatively long transmission line, and also to minimize distortion due to reflections.

Most interstage coupling networks are adjustable equivalent-T transformers with a passband amplitude response of 0.1 decibel over a 20-megacycle band centered at 70 megacycles. Exceptions are the coupling between the fourth and fifth stages and the output transformer, where fixed coupled transformers are used.

Three methods are used to minimize variations in amplifier gain from tube aging and supply voltage fluctuations. These are: (A) Regulation

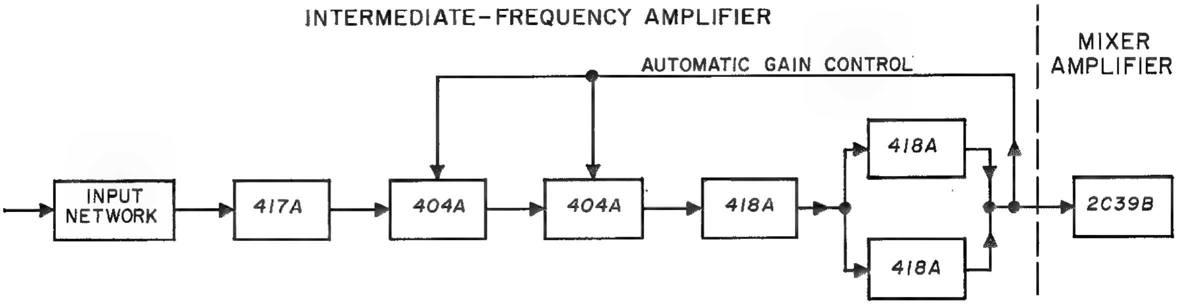

Figure 6—Block diagram of intermediate-frequency amplifier in converter–amplifier.

of the filament power supplies by a 12-volt filament supply and appropriate dropping resistors. (B) Positive grid bias and large cathode resistors. (C) Automatic-gain-control feedback from the intermediate-frequency output to the grids of the second and third stages.

1.5 Mixer–Amplifier

The mixer–amplifier consists of three *2C39B*'s in double-tuned radial transmission line cavity-coupling networks. The required bandwidth is obtained by using stagger-damped interstage coupling and single-tuned broad-band output coupling, giving an over-all 3-peaked response having a peak-to-valley ratio of approximately $\frac{1}{4}$ decibel over a 20-megacycle bandwidth.

The mixer–amplifier, Figure 4, consists of a mixer, two ultra-high-frequency amplifiers, and radial transmission line cavities. Each tube has one cathode cavity and one plate cavity. Broad bandwidth is obtained by adjusting the coupling between the plate cavity of one stage and the cathode cavity of the following stage. The local-oscillator power is coupled to the cathode of the mixer tube by a radial cavity, and the intermediate-frequency power is introduced through a broad-band lumped-constant network. The plate cavity of the mixer is loop-coupled to the cathode cavity of the first amplifier and proper coupling is obtained by the relative values of two tuning capacitors in the amplifier cathode cavity.

Figure 7—Front view, 10-kilowatt power amplifier.

A similar coupling arrangement is used be-

tween the first and second amplifiers. The second amplifier plate circuit is a single tuned cavity, loaded to the required Q by the 50-ohm load on the output loop. Capacitive-tuning plungers, inserted parallel to the axis, tune each cavity.

1.6 Power Supply

The power supply is nonregulated and supplies all direct and alternating voltages. Voltage regulation of ± 1 percent is obtained from an automatic line-voltage regulator mounted below the power supply.

A 60-second time delay between the application of filament power and plate power allows for filament warm-up. All high voltages are connected through relays on the primary side of the transformers, which are operated by the interlocks on the rear cabinet door and front dust covers.

2. *Power Amplifier*

The converter–amplifier output at approximately 10 watts is amplified to 10-kilowatt level by the power amplifier. The required 30 decibels of gain is obtained with a 6-cavity klystron (type *X631*, by Eitel-McCullough, Incorporated). This klystron and a set of tuned circuits covers the band from 680 to 900 megacycles.

2.1 Amplifier

Since the velocity modulation action at the gaps and the density modulation in the drift spaces are frequency-sensitive mechanisms, the klystron is inherently a narrow-band device. However, by utilizing broad-band tuned circuits at the klystron gaps, bandwidths of 3 to 4 percent of center frequency can be achieved.

The amplifier assembly is shown in Figure 7. The klystron assembly is on the left and the controls for the focusing-coil power supplies are on the upper right panel. The lower right-hand section houses the klystron filament transformer and the semiconductor rectifier for the focus supplies. Circuit breakers and overload indicators are aligned along the center panel and meters indicating the klystron parameters are in the top panel.

The klystron assembly, Figure 8, shows in greater detail the double-tuned circuits surrounding each klystron gap. The circuits across the first five klystron gaps consist of coaxial lines coupled to rectangular waveguide cavities. A waveguide section surrounds the klystron gap. To handle the high power level, the output circuit is formed by two waveguide cavities coupled together with an iris in the common wall between the guide sections. A lumped-circuit equivalent of these tuned circuits is shown in Figure 9.

The desired bandwidth is achieved by loading each circuit with an external resistor. In the input circuit, the waveguide section is loaded and the input signal is coupled to the coaxial section of the double-tuned assembly. Coupling and loading are adjusted to obtain maximum power transfer and maximal flatness of response (equal primary and secondary Q).[1] The interstage assemblies are loaded on one side only (the secondary coaxial section) for maximum impedance across the klystron gap at the given bandwidth.

The output primary tuned circuit is loaded by the klystron beam impedance; the secondary is loaded by the antenna impedance coupled through a capacitive probe. The loading resistors used in the first three tuned circuits are air-cooled. The fourth- and fifth-circuit resistors are $1\frac{5}{8}$-inch (4.13-centimeter) coaxial-line water-filled loads.

To simplify alignment of the amplifier, the six tuned circuits are synchronously aligned and centered about the carrier frequency. Each circuit is initially adjusted for transitional coupling with the klystron unenergized. A sweep signal generator is coupled to one circuit at a time and the response observed at the test point in the circuit. Then, under normal drive and power input conditions, a slight readjustment of the circuit parameters will result in the desired response. It should be noted that increased efficiency and gain can be obtained by "stagger damping" but at the expense of considerably greater adjustment difficulties.[2]

[1] M. J. Hellstrom, "Design Charts for Tuned Transformers," *Electronics*, volume 29, page 182; November, 1956.

[2] G. E. Valley, Jr. and H. Wallman, editors, "Vacuum Tube Amplifiers," McGraw-Hill Book Company, Incorporated, New York, New York, Massachusettes Institute of Technology Radiation Laboratory Series, volume 18; 1948: page 221.

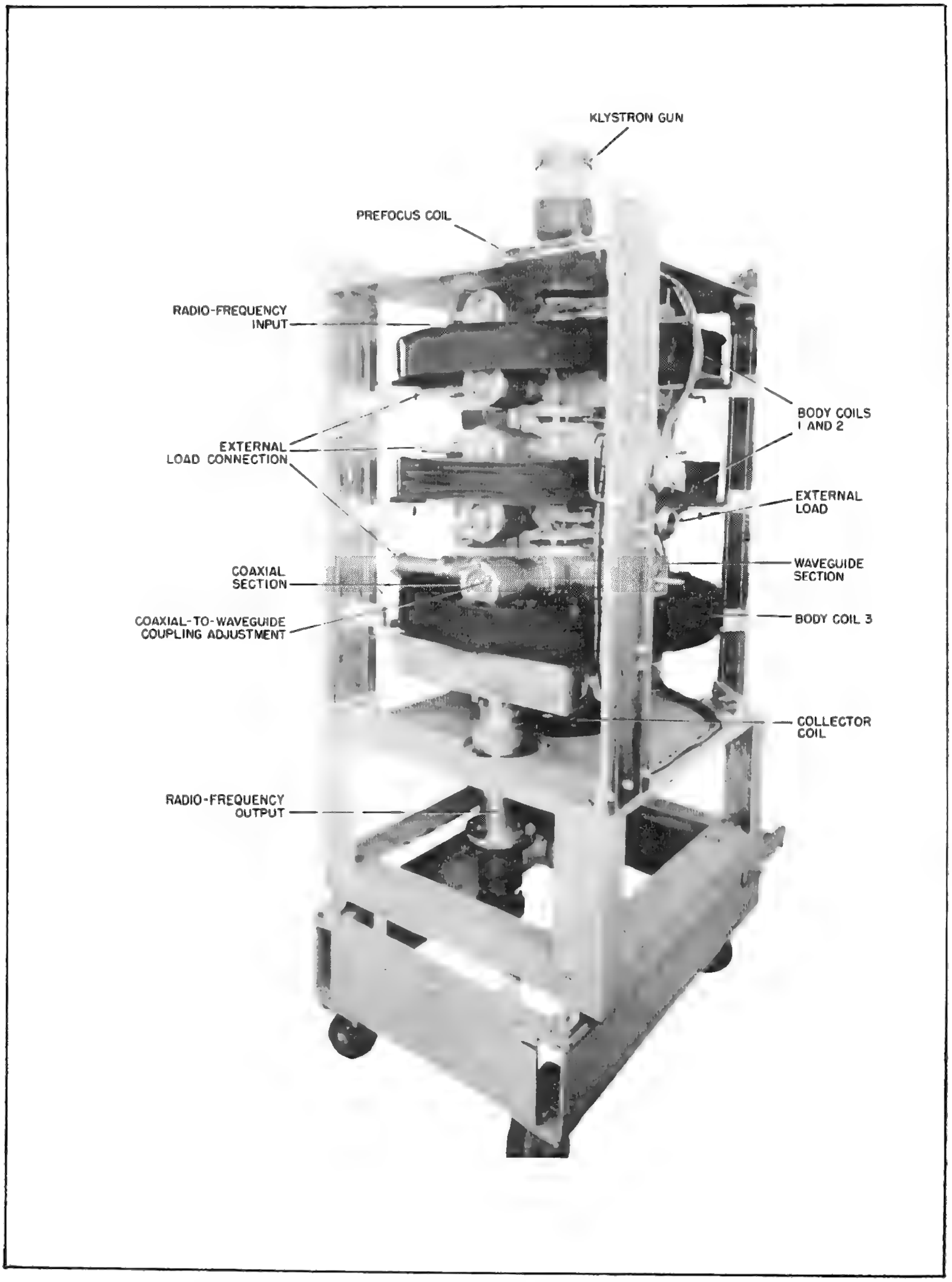

Figure 8—Klystron assembly and magnetic structure mounted on dolly.

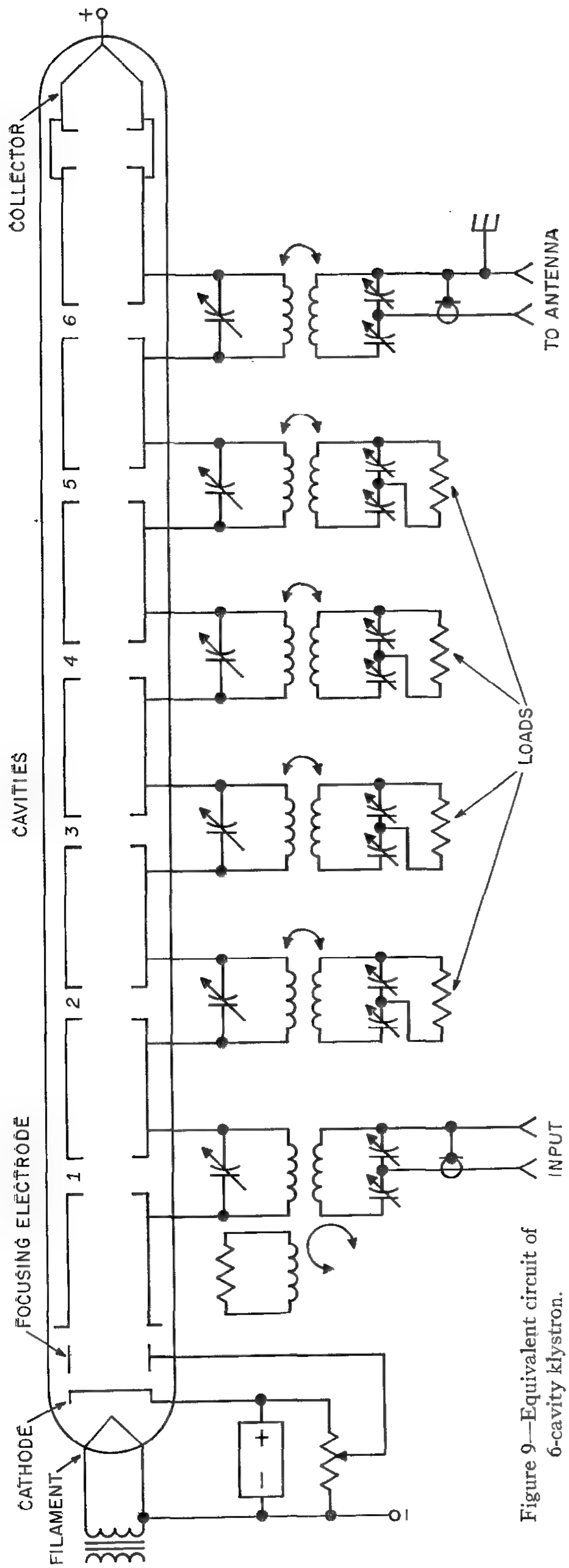

Figure 9—Equivalent circuit of 6-cavity klystron.

2.2 Power Supplies

The primary power to all filaments, to the klystron bombarder supply, and to the electromagnetic focusing supplies is regulated by an automatic motor-driven induction regulator to assure maximum tube life and minimum variation in klystron operating conditions.

All high-voltage magnetic components are nonflammable-oil-filled. The bombarder supply and beam supply are unitized assemblies; the magnetic components are contained in one oil-filled tank.

The bombarder supply utilizes xenon-filled rectifiers, whereas the beam supply has high-vacuum diodes to minimize the possibility of rectifier arcback. The circuits of both supplies are the 3-phase bridge type. The beam supply delivers 20 kilovolts at 2.5 amperes. Filtering of this unit is adequate to limit the incidental frequency modulation to less than 100 root-mean-square cycles per second. Stepless control of the beam voltage is possible at the amplifier control panel, over the range from 10.5 to 20 kilovolts at the nominal line voltage ± 10 percent.

2.3 Controls

Control and overload monitoring circuits provide proper sequential application of the various voltages and protect the equipment against progressive damage resulting from component failures or overloads.

All power-line circuits are monitored and switched by magnetic-trip circuit breakers with an appropriate tripping-time-versus-current characteristic that will accommodate starting current surges but also allow rapid tripping under large overload conditions.

High-speed direct-current relays monitor the following parameters: beam current, drift-tube current, fifth- and sixth-cavity voltage, excessive output transmission line mismatch, and low transmitter output. All these relays except the low-output relay will trip the beam power supply circuit breaker in less than 50 milliseconds. Pilot-light indications and an alarm call attention to a fault. In addition to electrical interlocks, the necessary mechanical interlocks such as liquid coolant flow, air flow, and high-voltage grounding switches are incorporated.

2.4 Cooling

Since the collector and drift tubes require liquid cooling, a closed system (coolant-to-air or coolant-to-water unit) is provided for each amplifier. Where an adequate supply of well water is available, the latter type of cooler is preferred. The air-cooled unit is shown in Figure 10 and the water-cooled unit in Figure 11. The coolant, a mixture of ethylene glycol and distilled water (60 to 40 parts per volume) with an acidic inhibitor, is also used as the dissipative element in the klystron fourth- and fifth-cavity loads.

2.5 Performance

Over-all transmission response of the converter–amplifier plus power amplifier is shown in Figure 12; the oscillogram shown was taken at 840 megacycles. Note that the 3-decibel bandwidth is 19 megacycles, and the 1.0-decibel points approximately 15 megacycles. Saturation power is 10.5 kilowatts at the upper end of the band (880 megacycles) and it falls to 8 kilowatts at the low end of the band (690 megacycles).

Gain at the level 0.5 decibel below saturation power is 32 decibels at the high-frequency end of the band and 28 decibels at the low end for the bandwidths shown. For smaller bandwidths, the gain can be increased to as much as 48 to 50 decibels before instability problems arise.

Envelope delay versus input frequency of the system, including the wide-band terminal equipment, converter–amplifier, power amplifier, and receiver is shown in Figure 13. The envelope delay curve is resolved into three components to obtain a standard basis of comparison; the slope component is found by drawing a straight line between the frequency points of interest, such as 66 megacycles and 74 megacycles of Figure 13; the difference in time delay noted between the two points is the slope. The change in envelope delay remaining after subtracting the slope component is the parabolic component, provided ripple is negligible.

If appreciable ripple is present, this must be averaged about the slope component and subtracted from the total envelope delay; the remainder is the parabolic component. Thus, the

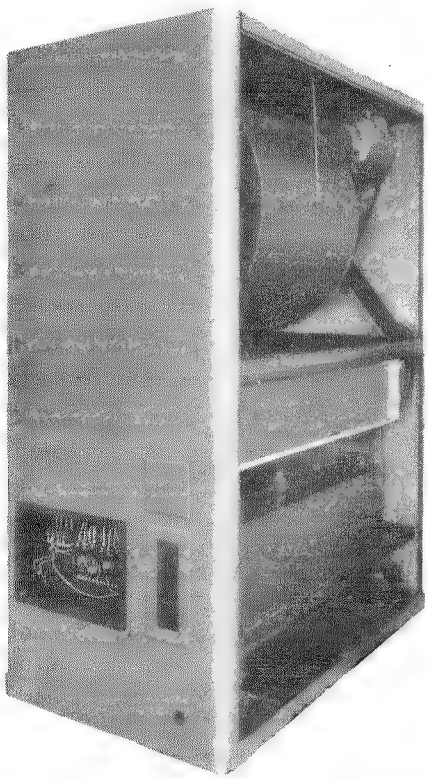

Figure 10—Heat exchanger for coolant to air.

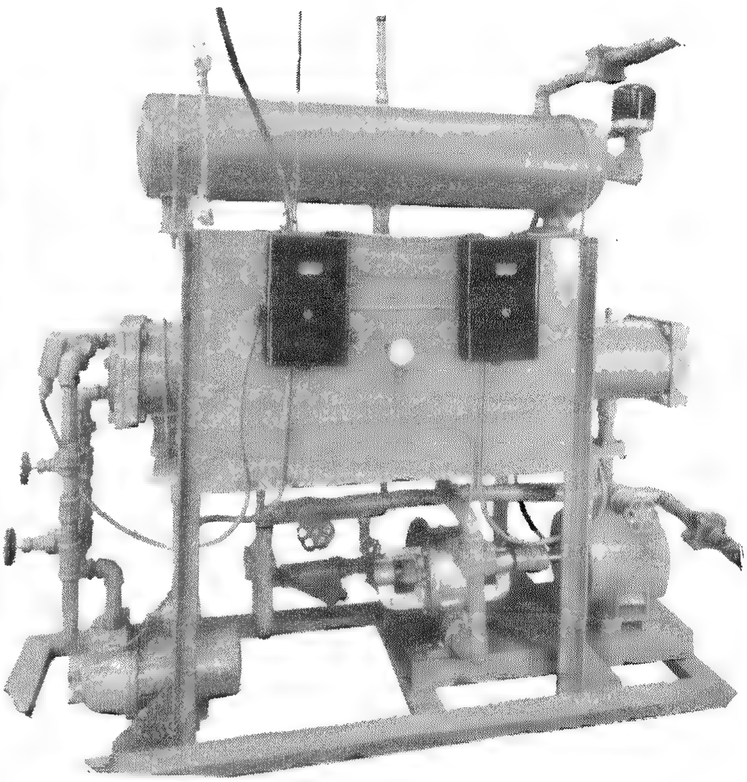

Figure 11—Heat exchanger for coolant to water.

sum of the three components at each frequency should equal the recorded curve. The components of time delay are given in Table 3.

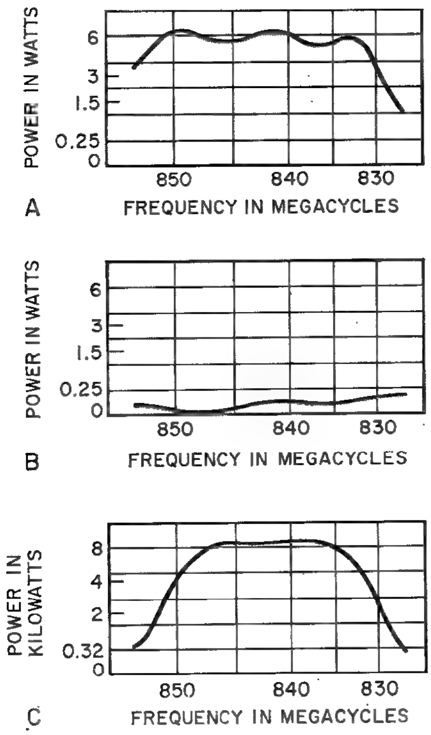

Figure 12—Amplitude response versus frequency. *A* = converter–amplifier. *B* = reflected from klystron input. *C* = 10-kilowatt amplifier driven by converter–amplifier.

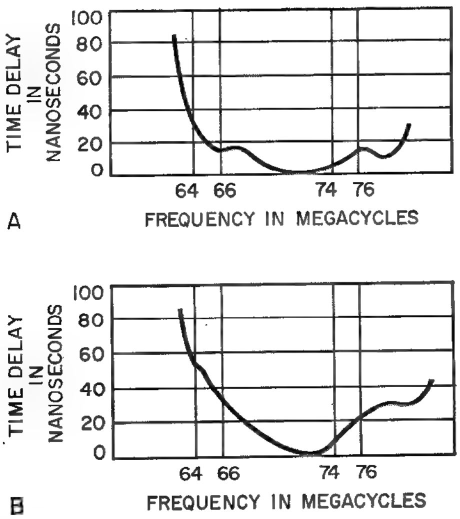

Figure 13—Envelope delay versus frequency. *A* = modulator, demodulator, and converter–amplifier. *B* = modulator, demodulator, converter–amplifier, and 10-kilowatt power amplifier.

Over-all response from intermediate-frequency input to the converter–amplifier to intermediate-frequency output at the receiver is shown in Figure 14 and the video-to-video response from the transmitting terminal through the radio system and the receiving terminal is shown in

TABLE 3

Time Delay Components

Component	Time Delay in Nanoseconds	
	At ±4 Megacycles	At ±6 Megacycles
Parabolic	15	22
Slope	25	30
Ripple	±2	±2

Figure 15. Intermodulation distortion is 50 decibels down, as measured through the system with a 500-kilocycle-bandwidth noise source producing a root-mean-square deviation of 0.4 megacycle.

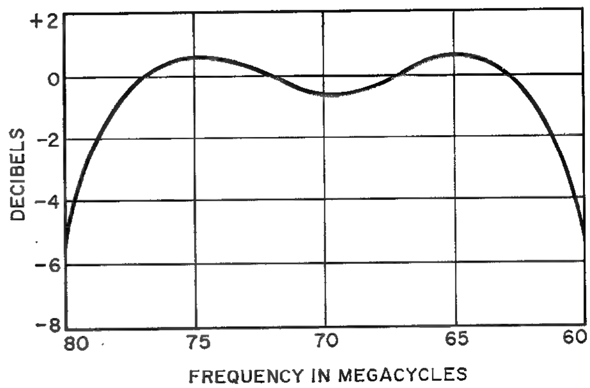

Figure 14—Over-all amplitude response versus frequency through converter–amplifier, power amplifier, and receiver.

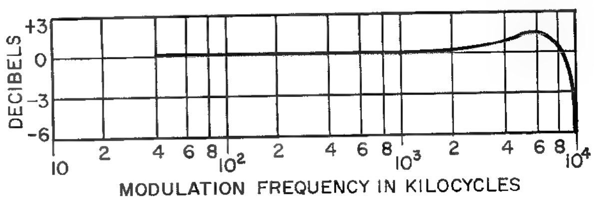

Figure 15—Over-all video-to-video amplitude response, including wide-band modulator and demodulator.

3. *Antenna System*

The antenna system consists of two 60-foot (18.3-meter) paraboloidal reflectors at each site, fed by dual-polarization horns, each of which is connected to one transmitter and two receivers. The transmission line connecting the transmitters to the antennas is *WR-1150* waveguide, and the receiver connection is by $3\frac{1}{8}$-inch (7.94-centimeter) 50-ohm Styroflex cable.

3.1 Reflectors

The reflectors and horn-support structures are shown in Figure 16.

3.2 Transmission Lines

WR-1150 copper-clad steel waveguide is employed for the high-power connection between the horn and the transmitter. The waveguide runs include both rigid bends and flexible waveguide bends at appropriate points on the tower and horizontal lines. Waveguide is used because of its low loss and its high power-handling capacity.

The receiver connection is made through $3\frac{1}{8}$-inch (7.94-centimeter), 50-ohm polyethylene-jacketed Styroflex cable. This large cable is used primarily because of its low loss and lower cost compared with waveguide.

3.3 Horn

The use of dual orthogonally polarized horns (greater than 50-decibel decoupling between polarizations) permits a wide variety of diversity system arrangements to be employed. In particular, the polarizations are selected in such a way that no high-power diplexers are required in the transmitter lines. This is accomplished by transmitting one frequency and receiving two frequencies at each horn. Received signals are separated by the receiver preselecting filters arranged as a diplexer.

The dual polarized horns for the 60-foot (18.3-meter) paraboloidal reflectors were designed to direct a signal at the reflector edge at approximately -10 to -12 decibels relative to the signal directed toward the reflector vertex.[3] Typical data are given in Figure 17. Since the horns were designed for a specific application, no particular effort was directed toward broad-band development.

Figure 16—On the facing page is shown a 60-foot (18.3-meter) paraboloidal antenna at Guanabo, Cuba.

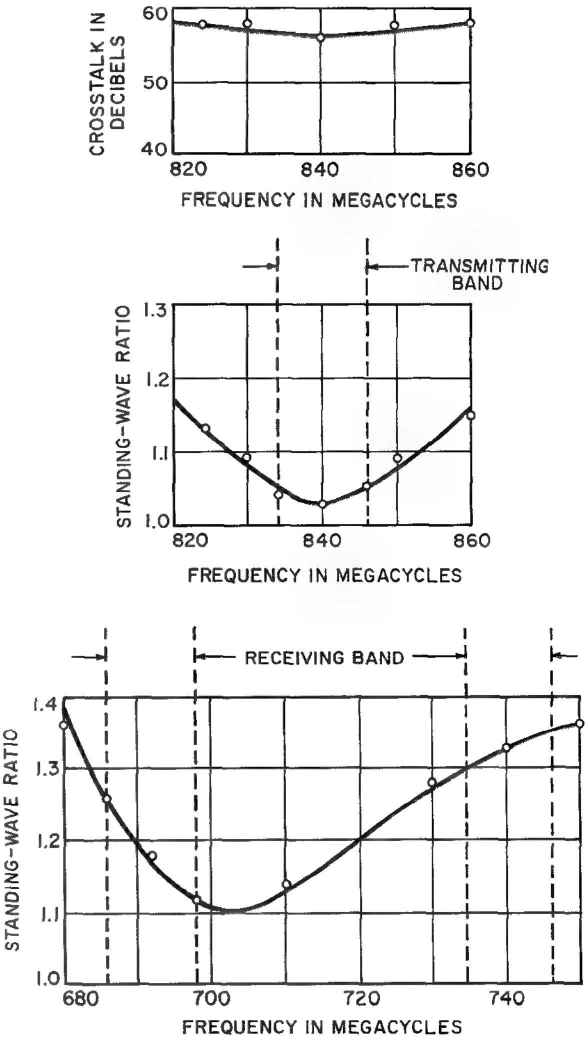

Figure 17—Standing-wave ratio and crosstalk of a typical dual-polarization horn.

An outline drawing of the horn is shown in Figure 18. Short capacitive posts match the horn at the individual transmitter frequency bands. This was necessary to obtain the required low voltage standing-wave ratio, less than 1.08, over the transmitter band. The receiver connections

[3] D. J. LeVine and W. Sichak, "Dual-Mode Feed Horn for Microwave Multiplexing," *Electronics*, volume 27, page 162; September, 1954.

are matched by proper selection of probe length and diameter and by the fin position relative to the probes.

The horns have been satisfactorily tested with 10-kilowatt continuous-wave power applied to the waveguide connection. The transmission lines and horns operate with approximately 0.5 pound-per-square-inch (35-grams-per-square-centimeter) (gage) internal dry air pressure and have been tested with over 2 pounds (140 grams) internal pressure.

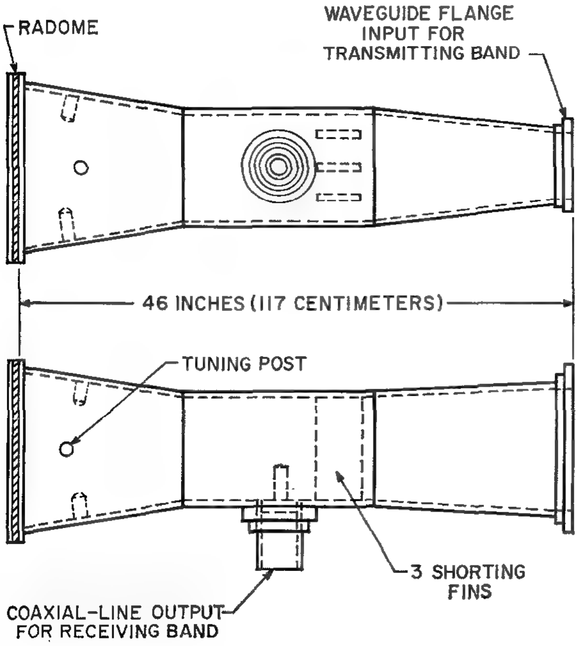

Figure 18—Outline drawing of dual-polarization horn.

The horn aperture barrier is a gasketed $\frac{1}{8}$-inch (3.2-millimeter) thick teflon-fiberglass sheet clamped to the aperture flange. Figure 19 shows a completed horn, ready for installation. The horn weighs approximately 150 pounds (68 kilograms). The maximum dimensions are approximately 14 by 14 by 46 inches (36 by 36 by 117 centimeters).

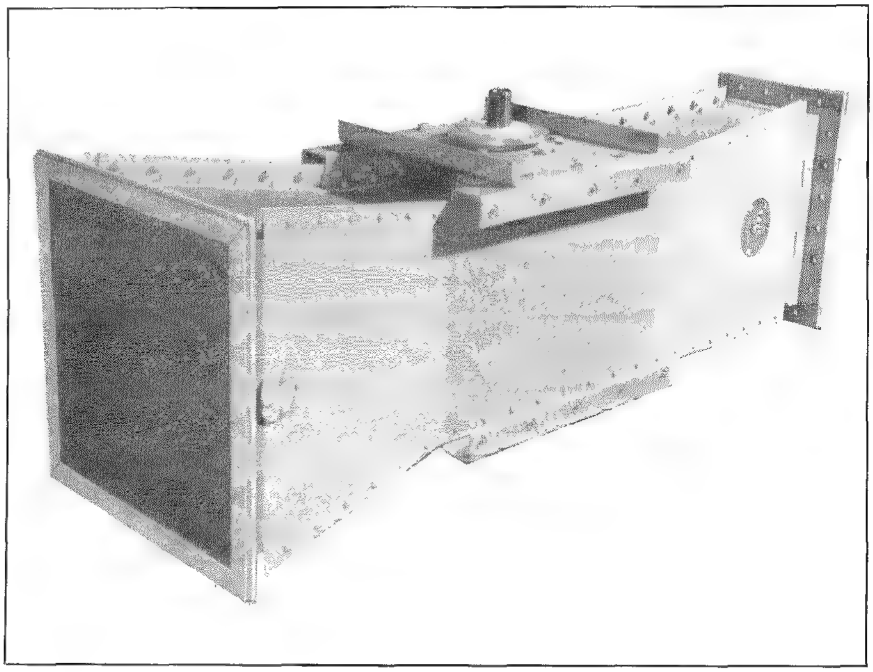
Figure 19—View of dual-polarization horn.

The antenna system performance can be described as follows: The free-space gain at midband is 41 decibels, not including line loss, and the midband half-power beamwidth is 1.4 degrees.

4. *Diversity Receiver*

The quadruple-diversity receiver set comprises two dual-diversity receiver sets operating at different frequencies in the range 680 to 900 megacycles with means for combining at the 70-megacycle intermediate frequency into one 70-megacycle output.

Figure 20 shows the receiver with dust covers removed. Two 9-foot (2.74-meter) racks (joined together at the center) hold the various 19-inch (48.3-centimeter) subchassis panels. Available components such as intermediate-frequency amplifiers, equalizers, and receiver control units, are used wherever possible.[4] Preselectors, mixers, local-oscillator–multipliers, combiners, diversity switching panel, and power distribution and alarm panels are designed specifically for this system.

[4] A. A. Roetken, K. D. Smith, and R. W. Friis, "TD-2 Microwave Radio Relay System," *Bell System Technical Journal*, volume 30, pages 1041–1077; October, 1951.

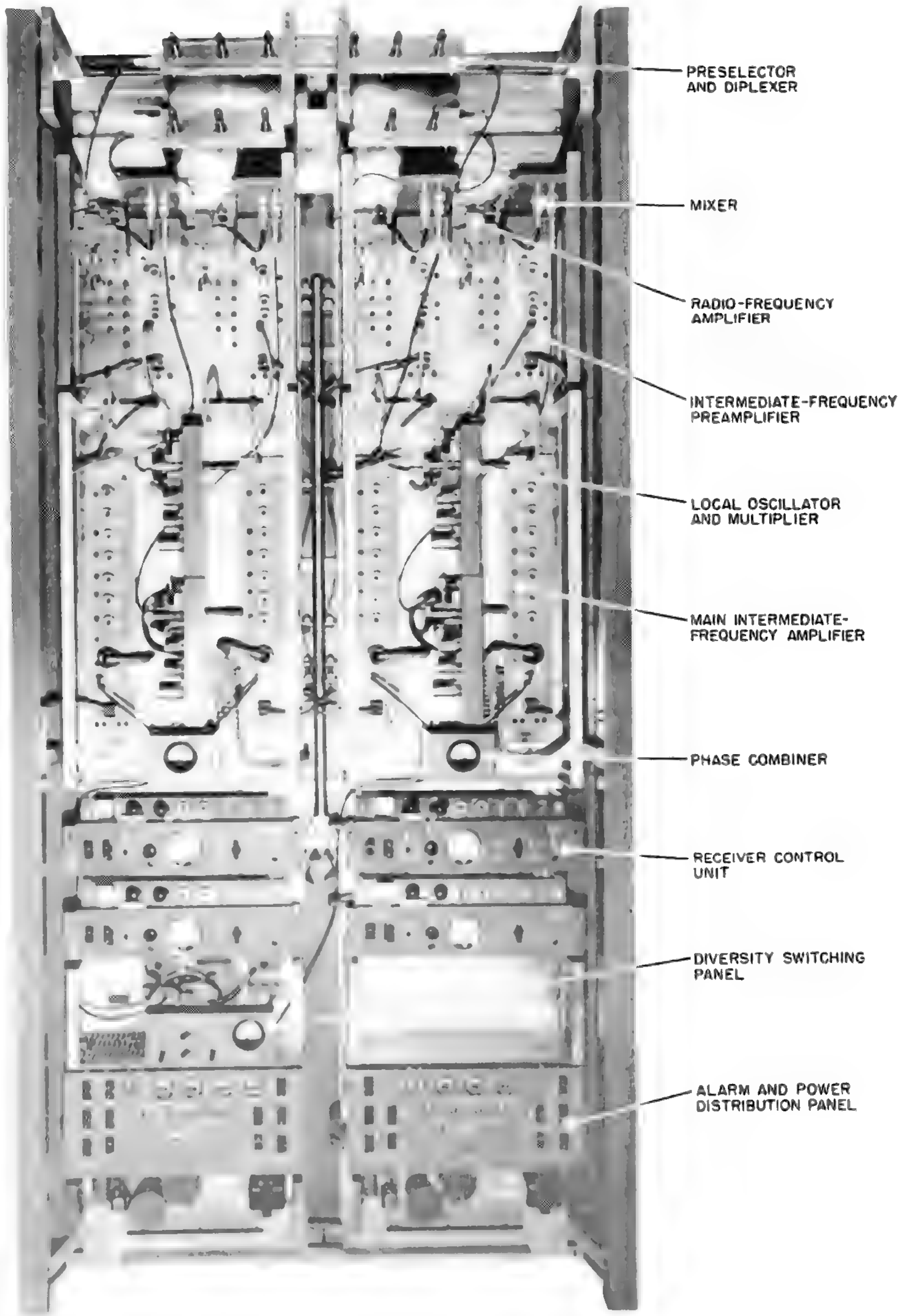

Figure 20—Front view of quadruple-diversity receiver.

Figure 21 is a simplified receiver block diagram. Preselectors separate the two frequencies from each antenna line for each of the four receiving channels. Each signal passes through a radio-frequency amplifier and mixes with the output of a crystal-controlled local-oscillator–multiplier to produce a 70-megacycle intermediate frequency. The intermediate-frequency preamplifiers, equalizers, attenuators, and main amplifiers deliver the proper level for phase combining. There are three phase combiners, one for each dual-diversity channel, and one to combine the dual-diversity outputs for quadruple combining. To meet the over-all delay distortion and amplitude response requirements, each major component of the receiver is so designed that the 0.1-decibel bandwidth is about 20 megacycles.

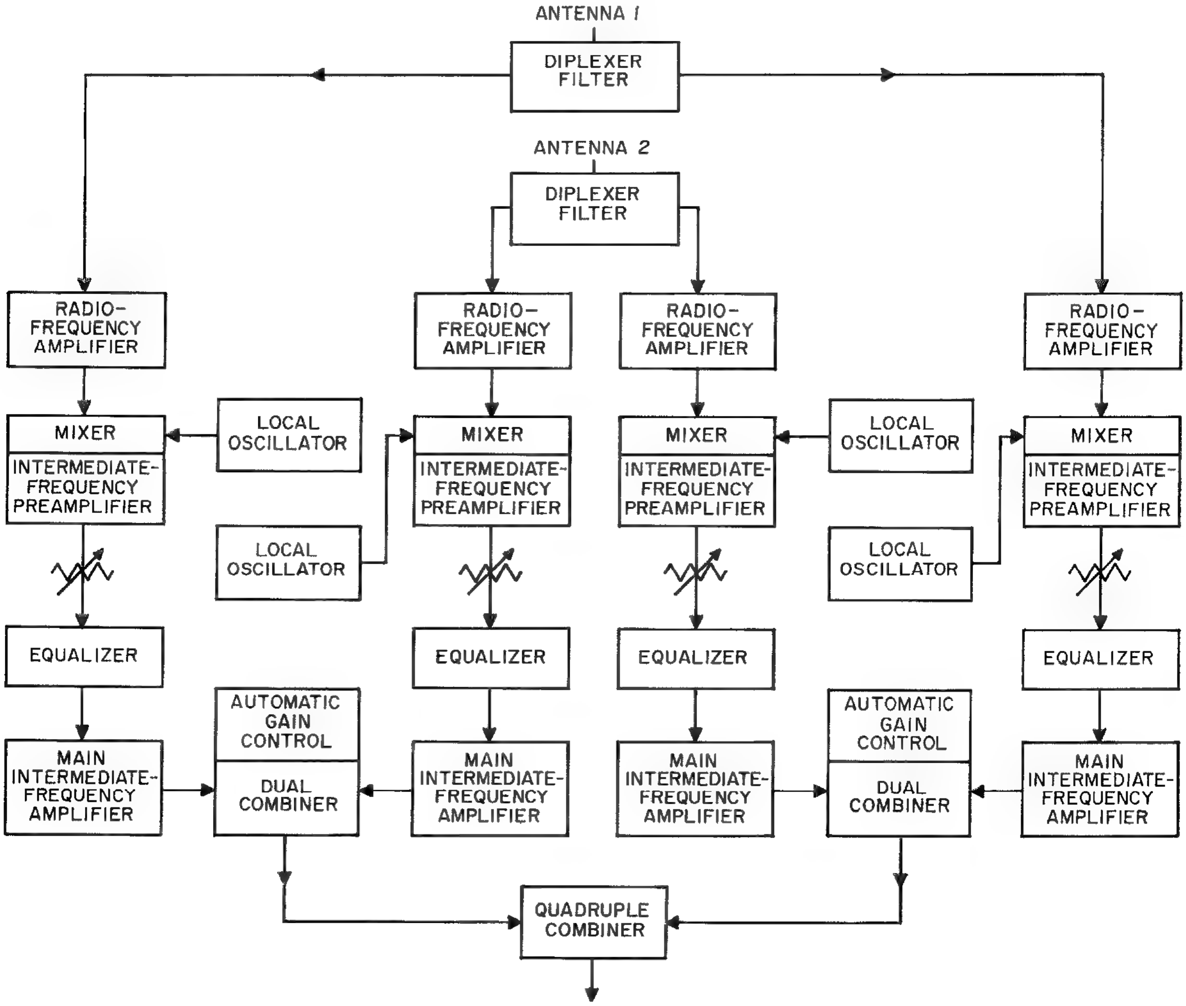

Figure 21—Block diagram of major receiver components.

4.1 Diplexer Filter

The diplexer is composed of two 3-cavity maximally flat amplitude response filters, centered at 692 and 740 megacycles, respectively, (or 840 and 880 megacycles), coupled to a T junction. The branch legs to each filter are a quarter wavelength at the center frequency of the other filter so that short circuits at the filter inputs are reflected as open circuits at the diplexer junctions. The correct leg is a matched line (standing-wave ratio under 1.1). The filter specifications are given in Table 4.

4.2 Radio-Frequency Amplifier

A single *416B* tube is used in a grounded-grid amplifier stage. Figure 22 is a top view. The

input standing-wave ratio is less than 1.3 over a 20-megacycle bandwidth. The plate circuit has two resonant transmission lines provided with adjustable coupling and an output impedance of

TABLE 4

RECEIVING FILTERS

	Frequency in Megacycles	Decibels Bandwidth	*Insertion Loss in Decibels*
Center Frequency	18.5	0.1	0.5
Other Received Frequency	30	1.0	30
Transmitter Frequencies	38	3.0	50
	175	40	

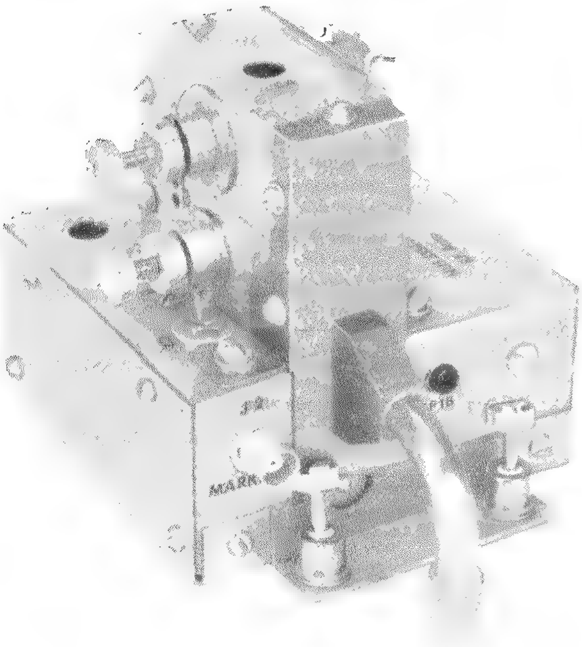

Figure 22—Low-noise radio-frequency amplifier.

50 ohms. Gains of 14 decibels and noise figures under 8 decibels are obtained.

4.3 MIXER PREAMPLIFIER

Microstrip construction is used in the balanced mixer.[5] The crystals (types *1N21B* and *1N21BR*) are each matched to a 50-ohm line by a short-circuited stub of proper length and position. Decoupling between the signal input and the local-oscillator arm is greater than 20 decibels.

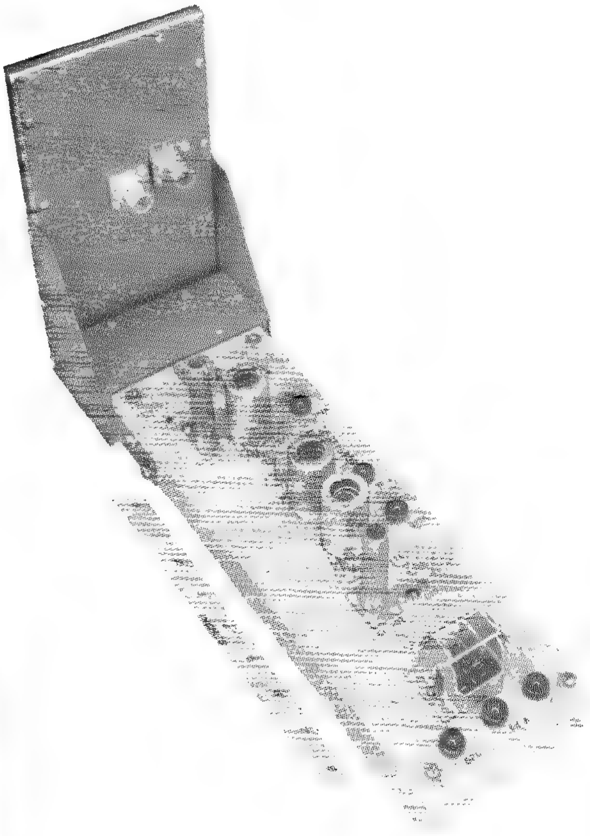

Figure 23—Microstrip mixer and 70-megacycle preamplifier.

The signal input standing-wave ratio is under 1.3 for a 20-megacycle bandwidth. The mixer is attached directly to the intermediate-frequency preamplifier, Figure 23, and drives the cathode of the first of three cascaded stages.

4.4 MAIN INTERMEDIATE-FREQUENCY AMPLIFIER

A wide-band intermediate-frequency amplifier, 8 tubes in cascade, provides 60-decibel gain at 70-megacycle center frequency.[4] The 3-decibel bandwidth is 32 megacycles and the 0.1-decibel bandwidth is 20 megacycles.

4.5 LOCAL OSCILLATOR–MULTIPLIER

The crystal-controlled local oscillator, Figure 24, uses a *12AT7* twin triode in a cathode-

[5] D. D. Grieg and H. F. Engelmann, "Microstrip—A New Transmission Technique for the Kilomegacycle Range," *Proceedings of the IRE*, volume 40, pages 1644–1663; December, 1952: also, *Electrical Communication*, volume 30, pages 26–35; March, 1953.

coupled circuit at frequencies in the 50-megacycle range. The output triode section is tuned to twice the crystal frequency. Two *404*-type tubes follow in cascade, the first as a 100-megacycle amplifier and the second as 300-megacycle tripler. This drives the grounded-grid *2C39* output tube either as a doubler or tripler in the frequency range 620 to 950 megacycles.

4.6 Phase Combiner at 70 Megacycles

Diversity reception is made possible by locking two 70-megacycle intermediate-frequency signals and adding them in phase. The phase detector supplies a direct control voltage to the local oscillator, which in turn adjusts their frequencies to produce phase lock. Except for two semiconductor diodes, the combiner uses only passive circuit elements. Part of the combined output voltage is used for automatic gain control and controls the main intermediate-frequency amplifiers. The 70-megacycle outputs of the dual combiner connect to the diversity switching panel.

4.7 Diversity Switching Panel

Outputs from the dual combiners are switched by coaxial relays to separate jacks or to a third combiner for quadruple operation. Automatic-gain-control voltages are simultaneously switched for either type of service. Figure 21 shows the system diagram for a quadruple receiver.

4.8 Power Distribution and Alarm Panels

Each dual-diversity receiver is separately powered by external lines from a commercial power supply. Incoming power is controlled by master switches, circuit breakers, and fuses. The alarm circuits monitor blower operation, radio-frequency signal level, and intermediate-frequency output; 3-phase lock conditions actuate an audible alarm whenever their operation is abnormal.

5. *Acknowledgments*

The successful development of this equipment is due to the teamwork and co-operation of a large number of people. The authors are grateful to the following engineers for their contributions: A. G. Kandoian, for the initial planning and over-all supervision; W. Sichak, for over-all technical supervision; J. Gulack for system layout; W. Glomb, H. E. Lott, and J. Simkovich for the converter–amplifier; L. Gray, H. Goldman, and W. Jacobus for the power amplifier;

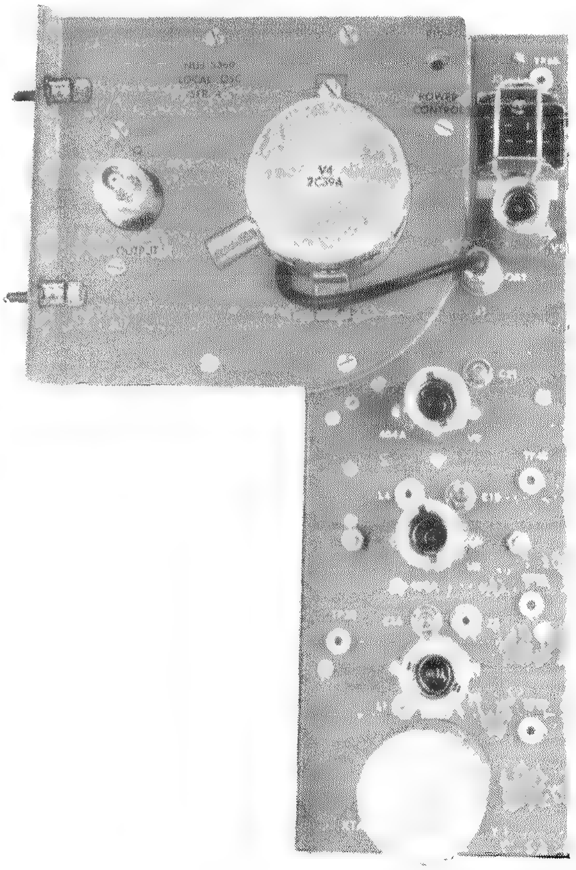

Figure 24—Crystal-controlled local oscillator.

R. T. Adams, A. T. Brown, and R. Bresk for the receiver; and L. Juhas for the feed horn.

The requirements and equipment specifications for the system were set up by a joint committee of representatives of Bell Telephone Laboratories, the Long Lines Department of American Telephone and Telegraph Company, Western Electric Company, International Telephone and Telegraph Corporation, and ITT Laboratories. Special acknowledgment is due to the following members of the committee: K. P. Stiles and A. A. Bottani of Long Lines Department, American Telephone and Telegraph Company; and H. E. Curtis, N. F. Schlaack, and H. A. Wells of Bell Telephone Laboratories.

United States Patents Issued to International Telephone and Telegraph System; May-July 1958

BETWEEN May 1, 1958 and July 31, 1958, The United States Patent Office issued 72 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

H. H. Abelew, Mackay Radio and Telegraph Company, Signal Selector Device, 2 834 003.

P. R. Adams and R. I. Colin, Federal Telecommunication Laboratories, Aircraft Radio Navigation System, 2 836 815.

R. T. Adams and R. E. Altoonian, Federal Telecommunication Laboratories, Microstrip Switch, 2 842 637.

P. R. R. Aigrain, Laboratoire Central de Telecommunications (Paris), Circuit Element Having a Negative Resistance, 2 843 765.

P. R. Aigrain, Laboratoire Central de Telecommunications (Paris), Semiconductor Crystal Rectifiers, 2 845 370.

D. F. Albanese, Federal Telecommunication Laboratories, Pulse Modulator 2 837 719.

J. L. Allison and B. Alexander, Federal Telecommunication Laboratories, Airborne Pictorial Navigation Computer, 2 836 816.

M. Arditi and J. Elefant, Federal Telecommunication Laboratories, Microwave Transmission Line, 2 833 995.

M. Arditi, Federal Telecommunication Laboratories, Traveling-Wave Electron Discharge Devices, 2 833 962.

R. J. Arndt, Kellogg Switchboard and Supply Company, Negative-Impedance Repeater Having Gain Controls, 2 844 669.

R. P. Arthur, Kellogg Switchboard and Supply Company, Electromagnetic Counting Device and Contact Bank, 2 844 686.

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E. De Faymoreau, Federal Telecommunication Laboratories, Relay Control Circuit, 2 833 967.

M. J. DiToro, W. Graham, and S. M. Schreiner, ITT Laboratories, Volume Expander, 2 845 599.

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R. H. Geiger, Federal Telecommunication Laboratories, Indirectly Heated Thermionic Cathode, 2 845 567.

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J. F. Houdek, Jr., Kellogg Switchboard and Supply Company, Telephone Substation Ringers, 2 844 767.

R. W. Hutton and E. J. Leonard, Kellogg Switchboard and Supply Company, Private-Branch-Exchange Line-Hunting System, 2 839 611.

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Methods of Recording and/or Modifying Electrical Intelligence

2 838 745

E. P. G. Wright and J. Rice

A data-processing system is described that includes a magnetic device on which information is stored in binary code on a plurality of tracks. The system provides an arrangement for reading the information from the storage device successively for the different items recorded, modifying this information in accordance with changes that should be made, and then re-recording the modified item in the same track as that from which it was originally read. The control of the system is effected by various control pulses operating through a system of coincidence gates and flip-flop circuits.

Omnirange-Beacon Antenna

2 836 820

S. Pickles, P. R. Adams, and C. Lucanera

An omnidirectional-beacon antenna for use in tacan has radiators in a vertical array to reduce the vertical angle of transmission. With each radiator, there are two sets of reflectors mounted on an insulating drum for modifying the pattern radiated from the antennas to produce a multi-lobe directive beacon pattern.

Aircraft Radio Navigation System

2 836 815

P. R. Adams and R. I. Colin

A radio beacon system producing a rotating directive pattern, such as is used in tacan, is covered. It provides for the transmission on this rotating pattern of reference and bearing signals, mobile-unit-identifying signal pulses, and message signal pulses together with controls for applying these various types of pulses in sequential order. The beacon can be used to cooperate simultaneously with a plurality of mobile units, such as aircraft.

Telephone Subscriber's Instruments

2 838 612

L. C. Pocock

A telephone subscriber's set is described that employs a transistor amplifier to permit the use of a resistive bridge network in place of an inductance coil for reducing side-tone effects. The transistor amplifier is associated with the resistive bridge network and serves to offset losses that would otherwise occur in this bridge network. The amplifier is in conjugate relationship with both the transmitter and receiver.

Composite Radiation Amplifier

2 837 660

R. K. Orthuber and L. R. Ullery

The radiation amplifier is of the type in which light or X-ray radiation images may be amplified by control of the voltage across an electroluminescent substance through the medium of photoconductive elements illuminated by the radiation forming the image. An arrangement is provided in which the electroluminescent particles and the photoconductive particles form a single composite layer either by a coating of photoconductive material on the electroluminescent particles or by intermixing of the two substances so that the photoconductive material is effectively in shunt with the electroluminescent particles.

Circuit Element Having a Negative Resistance

2 843 765

P. R. R. Aigrain

An electric circuit taking advantage of the properties of semiconductor diodes, such as germanium diodes, was developed to provide a circuit element having a negative-resistance characteristic. A direct-current source and an alternating-current source are applied in series across a semiconductor rectifier. The alternating-voltage is of higher amplitude than the difference in voltage between the terminals of the direct-current source. The frequency of this alternating current is chosen so that the irregular carriers of electric charge injected into the semiconductor at a particular point in the cycle of the alternating voltage do not disappear completely during one period of the alternation.

Contributors to This Issue

Johann Augustin

Johann Augustin was born in Berlin, Germany, on October 8, 1909. After practical training as a mechanic, he entered a public technical college from which he received an engineering degree in 1935.

Mr. Augustin has been employed by Standard Elektrik Lorenz since 1935. Prior to the second world war, he was a designer in the Berlin plant. After the war, he was placed in charge of a group concerned with the development of teleprinters including page and tape printers as well as tape perforators and readers. He is author of three papers in this issue on such equipment.

A. H. Beck

A. H. Beck was born in Norfolk, England, in 1916. He received a bachelor of science degree in engineering from University College, London.

After two years with Henry Hughes and Sons, he was assigned to the Admiralty Signals Establishment until the end of the war. In 1947, he left Hughes to establish the Enfield valve laboratory of Standard Telephones and Cables, which later became the valve division of Standard Telecommunication Laboratories. He is coauthor of the paper in this issue on gas tubes. In 1958, Mr. Beck joined the academic staff of Cambridge University as a lecturer in engineering. He is a member of Queens' College.

Mr. Beck is an Associate Member of the Institution of Electrical Engineers and a Fellow of the Institute of Radio Engineers. He is the author of three books on electronics, "Velocity Modulated Thermionic Tubes", "Thermionic Valves", and "Space-Charge Waves".

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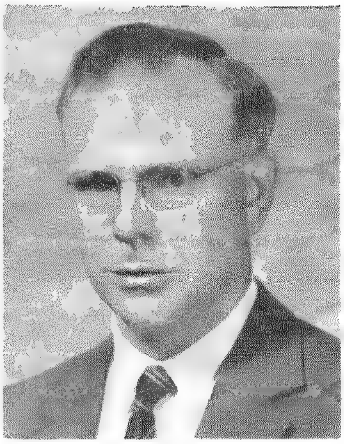

W. F. S. Chittleburgh

William F. S. Chittleburgh was born in London on October 2, 1923. He studied electrical-communication engineering at the City and Guilds College, London, where he received in 1943 an Associateship of the City and Guilds Institute together with a bachelor of science (engineering) degree.

On graduation, he joined Standard Telephones and Cables, and was initially associated with voice-frequency telegraphs. In 1948, after a period of working on coil and transformer design, he returned to the voice-frequency telegraph and signalling section, where he is at present in charge of such developments. He is the author of an article on voice-frequency telegraph equipment in this issue.

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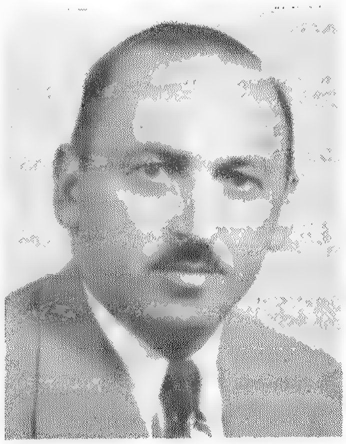

Robert A. Felsenheld

Robert A. Felsenheld was born on February 15, 1910, in East Orange, New Jersey.

He has been active in the broadcast field since 1927. In 1941, he joined ITT Laboratories and is now a senior project engineer in the radio communication laboratory. He has been active in the development of very- and ultra-high-frequency components, particularly antennas, transmission lines, and receivers. He reports in this issue on some over-the-horizon equipment.

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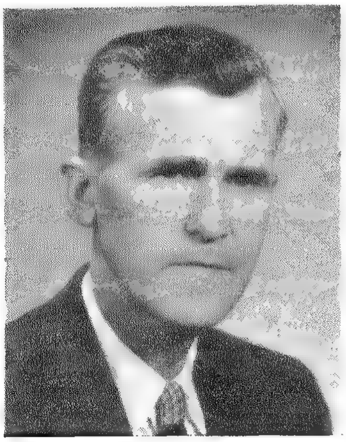
H. Havstad was born on August 15, 1907. He received the degree of bachelor of science in electrical engineering from the College in Oslo, Norway, in 1931.



H. Havstad

Mr. Havstad was the radio engineer for the Veslekari north-pole expedition in 1928. From 1931 to 1939, he did radio field engineering, installation, and operation for Tropical Radio Company and several merchant marine organizations.

He did research and development work at Bell Telephone Laboratories from 1939 to 1946. During the next two years, he was self-employed.

Mr. Havstad came to ITT Laboratories in 1948. His main line of work has been on pulse-modulation microwave links and on beyond-the-horizon systems. In this issue, he is coauthor of a paper in the latter field.

He is a member of the Institute of Radio Engineers.

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J. D. Holland. A photograph and biography of Mr. Holland, author of the paper on transition-rate discriminators, appears in the September 1959 issue of *Electrical Communication.*

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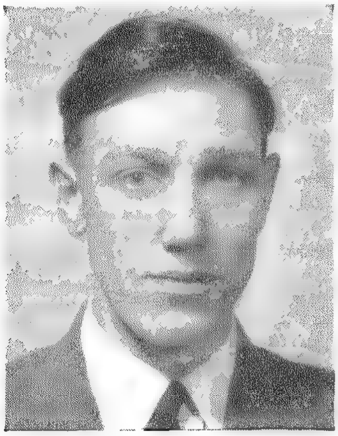

T. M. Jackson

T. M. Jackson was born in Merioneth, Wales in 1921.

He joined the Royal Air Force in 1940 and spent four years in experimental work on radar systems at the Government Telecommunication Research Establishment.

In 1946, he joined the Enfield valve laboratory of Standard Telephones and Cables and was transferred to Standard Telecommunication Laboratories in 1954. For several years he has been concerned with gas discharge phenomena and has been responsible for the development of special gas discharge devices, on which subject he is coauthor of a paper in this issue. Mr. Jackson is now doing microwave work, particularly on the problems of power generation at millimetric wavelengths.

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Bent Bulow Jacobsen

Bent Bulow Jacobsen was born in Kolding, Denmark, in 1906. He graduated in 1928 from the City and Guilds of London Engineering College.

He joined the International Telephone and Telegraph Laboratories, London, in 1929 and later transferred to Standard Telephones and Cables Limited. He has worked for many years on carrier system design and more recently also on microwave radio links. He has speciallized in the study of noise in long-distance transmission. Mr. Jacobsen is the author of the paper in this issue on probability theory in telephone transmission.

He is a Member of the Institution of Electrical Engineers and of the Institution of Danish Civil Engineers.

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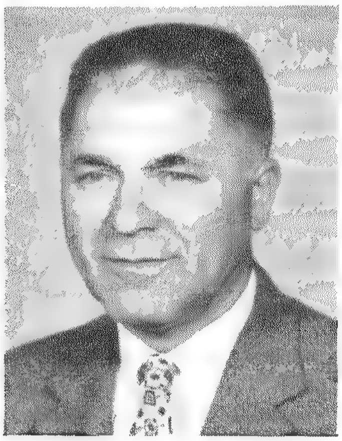

Jack L. Jatlow

Jack L. Jatlow was born on April 7, 1902. He received a bachelor of science degree in electrical engineering from Rensselaer Polytechnic Institute in 1924.

From 1924 to 1931, he was associated with the Conner Crouse Corporation. From 1931 to 1932, he was with Wired Radio Corporation, and for the following three years served as assistant chief engineer of F.A.D.Andrea Radio Corporation. From 1935 to 1940, he was employed by Photo Positive Corporation for research on photographic emulsions. He was chief engineer of Republic Engineering Products from 1941 to 1942.

In 1942, he joined Federal Telephone and Radio Corporation. Since his transfer to ITT Laboratories in 1954, Mr. Jatlow's main duties have been those of project manager for various beyond-the-horizon radio link systems. In this issue, he is coauthor of a paper reporting on one of these systems.

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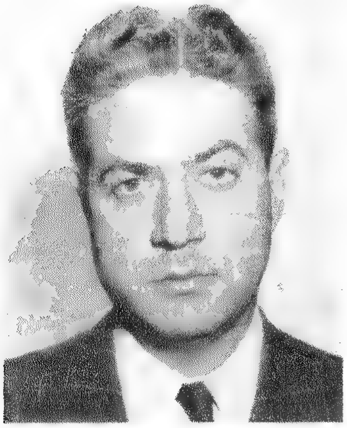
Donald J. LeVine was born in Brooklyn, New York, on October 10, 1921. He received two degrees in electrical engineering, the bachelor of science from the College of the City of



Donald J. LeVine

New York in 1943 and the master of science from the Polytechnic Institute of Brooklyn in 1952.

From 1943 to 1946, he served in the Army Signal Corps working with radar and pulse-time-modulation microwave relay communication equipment. From 1946 to 1948, he was with the Microwave Research Institute of the Polytechnic Institute of Brooklyn, with a primary interest in microwave power-measuring devices.

In 1948, Mr. LeVine joined ITT Laboratories, where he has been active in the development of microwave systems, components, and antennas. He is coauthor of the paper in this issue on beyond-the-horizon radio equipment.

Mr. LeVine is a Senior Member of the Institute of Radio Engineers, an Associate of the American Institute of Electrical Engineers, and a registered professional engineer in the state of New York.

Alexander William Montgomery

• • •

Alexander William Montgomery received the B.Sc. Tech. degree from Manchester College of Technology after having served for four years with the army during World War I.

In 1921, he joined the Western Electric Company in London and became closely associated with the early development in Great Britain of repeaters and carrier systems. He was active also in the field of voice-frequency telegraphy. He is now joint general manager and a director of Standard Telephones and Cables, Limited, and vice-chairman of Standard Telecommunication Laboratories, Limited. During World War II, he was actively engaged on the development of the British defense teleprinter network and of a wide range of communication equipment. He served on a number of government committees and, in 1944, visited the U.S.A. as a member of a Ministry of Supply mission. For his war work, he was made an Officer of the Order of the British Empire.

Mr. Montgomery represents the British telecommunications industry on a number of governmental and other committees.

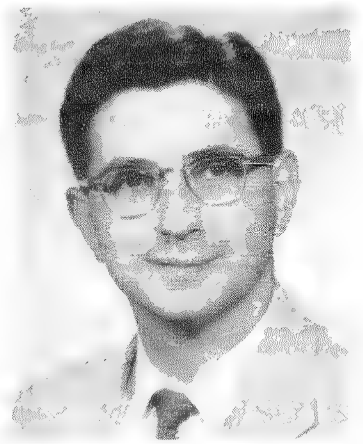
Mr. Montgomery reports in this issue on the past and future of coaxial-cable communication systems.

He is a Member of the Institution of Electrical Engineers and of the American Institute of Electrical Engineers, and a Fellow of the Institute of Radio Engineers.

• • •

L. Pollack was born in New York City on November 4, 1920. After receiving his bachelor of science degree in electrical engineering from the College of the City of New York in 1941, he worked at Fort Monmouth Signal Development Laboratory and in the Alaska Defence Command on the installation and maintenance of radar and communications equiment.

In 1943, Mr. Pollack joined ITT Laboratories, where he is now an executive engineer. His main course of work has been in the design of very- and ultra-high-frequency receivers and high-power transmitters. He is a coauthor of the article in this issue on beyond-the-horizon radio equipment.



L. Pollack

Mr. Pollack is a Senior Member of the Institute of Radio Engineers. He is a member of the relay committee and of the television transmitter committee of the Electronic Industries Association.

• • •

Werner Schiebeler was born in Bremen, Germany, on March 17, 1923. He received a diploma in physics in 1952, from Göttingen University. Three years later, he was awarded a doctorate in natural sciences from the Max Planck Institute in Göttingen.

Dr. Schiebeler has been on the staff of Standard Elektrik Lorenz in Pforzheim since 1955 and is employed in the development of electronic teleprinters and electronic input and output devices for data-handling systems. He reports in this issue on a teleprinter synchronizign set.

Werner Schiebeler

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